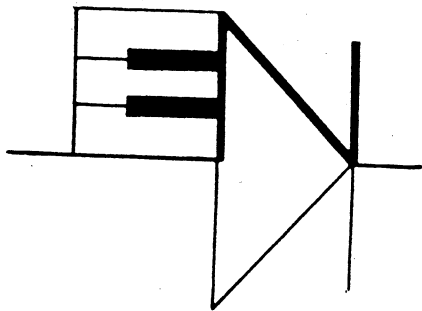




# ELECTRONOTES

## BUILDER'S GUIDE and PREFERRED CIRCUITS COLLECTION

2010

~~2008~~

2005

~~2000~~

This ~~1995~~ printing of the EBG&PCC is the same as the EBG&PCC, 1985-1986 edition, July 1986. No changes have been made.\*

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Electronotes, Newsletter of the Musical Engineering Group  
Musical Engineer's Handbook  
Electronotes Applications Notes

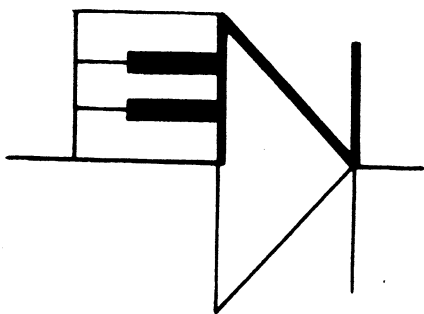
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\* slightly rearranged for this printing







# **ELECTRONOTES**

## BUILDER'S GUIDE AND PREFERRED CIRCUITS COLLECTION

- ▷ PUBLISHED BY: ELECTRONOTES  
1 PHEASANT LANE  
ITHACA, NY 14850-6399  
(607)-273-8030
- ▷ COST: \$15.00 (NON-SUBSCRIBERS TO ELECTRONOTES NEWSLETTER)  
\$10.00 (CURRENT SUBSCRIBERS TO ELECTRONOTES NEWSLETTER)

▷ PUBLISHING, ORDERING, AND SHIPPING INFORMATION:

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ELECTRONOTES BUILDER'S GUIDE  
AND  
PREFERRED CIRCUITS COLLECTION

PART 1: INTRODUCTION, BASICS, PLANS

PART 2: CONSTRUCTION PRACTICES

PART 3: PREFERRED CIRCUITS COLLECTION

PART 4: APPENDICES



# PART ONE

## INTRODUCTION, BASICS, PLANS

- 1-1 INTRODUCTION
- 1-2 PLAN OF ATTACK
- 1-3 THE ACTUAL PLAN OF AN  
ELECTRONIC MUSIC SYSTEM
- 1-4 DECIDING WHAT WILL BE IN YOUR SYSTEM
- 1-5 HOW TO OBTAIN THE PARTS YOU NEED
- 1-6 BUILDING YOUR SYSTEM
- 1-7 SYSTEM INTEGRATION
- 1-8 STANDARDS
- 1-9 SUBSTITUTION OF OP-AMPS
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## INTRODUCTION:

This builder's guide is basically a collection of circuits and building information selected from Electronotes and related publications. We have tried to present this information at a practical and reasonable level. We have had to assume that the reader and user of this guide has some experience in electronic construction. It is the nature of the complexity of useful electronic music systems that a "zero based" approach would involve too much description and would leave many readers "over their heads" at various points along the way. We believe this guide can be successful only if the user does end up with a useful system in a reasonable length of time.

Let us spend a few moments discussing some things that we believe to be true as a result of our observations over the years. First, we believe that a dedicated and resourceful builder can construct a useful electronic music system at a cost of approximately 1/6 that of a corresponding commercial unit. Secondly, to achieve this goal in a reasonable length of time, it is necessary to cut a few corners with regard to internal appearance. That is, the inside of your unit may not look professional, but it will be electrically and mechanically sound just the same. Thirdly, the amount of effort that goes into a fair sized unit is quite large (many many evenings) so unless building is part of the "fun", you may be better off buying a commercial unit, at least to get started. Finally, in a fair sized unit, something is almost certain to go wrong somewhere along the line so you must be prepared for some frustration and hunting of "bugs".

As an initial indication, glance ahead at some of the schematics and ask yourself if you could build and troubleshoot the circuits. We are not concerned with the efficiency of your efforts, but could you do it? If you feel that you could work from these schematics, locate the parts, assemble the unit, and get it working, then you can use this guide. The good news is that much of the other information in this guide is chosen to make your efforts much easier. Some of the things that you could have done (with great difficulty) will turn out much easier after you follow our suggestions.

Construction of an electronic music system naturally falls into several phases: planning, parts acquisition, actual building, testing and troubleshooting, and finally, system integration. Overall organization of your efforts during each phase cannot be stressed to much. The planning phase is really the most important. Yet the most important thing to realize about the planning phase is that it must end. The planning phase must end when a "clear picture" of what is to be done overall emerges. Details along the way will naturally upset plans carried out in detail, so don't waste your time. Parts acquisition is not difficult at all (assuming you have the funds at hand). Actual building is not a problem once you get going. Staying going is another problem that can be helped if you have organized well to begin with (avoid being held up for lack of a minor but essential tool, for example). Testing and troubleshooting are difficult to predict. Fortunately, most everything will work properly if constructed properly with quality parts. Yet, expect at least several circuits to give you some trouble for various reasons. System integration is the most fun. This is where you put things together and make sub-systems into a single unit. You will have no problems here unless you failed to grasp the "clear picture" during the planning phase.

At this point, it would be useful to glance through the entire guide to get a feeling for what is included. Then you can go on to the "plan of attack" that follows and work from there.



## **1-2** PLAN OF ATTACK:

A plan of attack is not a plan for building an electronic music system. It is a plan for seeing clearly all that is involved. In brief, it involves looking over this builder's guide carefully, considering your personal resources carefully (your technical ability, your funds, and your personal "drive"), and then, assuming you decide to go ahead, reflecting on the entire situation.

After you have gone over your basic resources, the next step is to build the whole device in your mind. This may seem like a complete waste of time, but it is a very useful step, first for pointing out problem areas, and secondly for giving you building experience (albeit simulated experience). Build it in your mind going into each aspect of construction in detail for at least one example. That is, you need not imagine drilling every hole in a panel, but you should imagine drilling one hole in detail. Such details could include such things as "I need to get a center punch at the store" and "Which neighbor is it who borrowed my electric drill?" and "I have to measure two pots side by side to see how close to drill the holes" and so on. Take notes on things you have to do and things you have to find out.

It is important to realize that during this plan of attack, it is not necessary to find the answers to all your questions. A note such as "I have to find out how to heat sink power supply regulators" will serve at this point in lieu of actually finding out. What we are looking for is a plan of attack that is complete in its scope, not necessarily in its details. You must mentally build your equipment from schematic diagram to turning on the power switch. Don't avoid or neglect anything. Nothing will happen by magic. You are responsible for making things happen.

Once you have mentally build your device, even with fragmentary details, you have completed your plan of attack except for possible revisions of the order in which things are to be done. You know what is to be done, how to do much of it, and in a sense, have done it once in a "dry run." If you are not satisfied at this point, try to find more answers in this guide, by asking friends questions, or give us a call at the number given in this guide.

## THE ACTUAL PLAN OF THE ELECTRONIC MUSIC SYSTEM:

If you have completed your "Plan of Attack", you are now ready to go ahead with the actual plan of the system. There are three areas to cover here. The first is the plan of the actual system - what are you going to include in your system, and which circuits are you going to choose. Here we should point out right away that while this system plan may seem a formidable task, since you may have little idea about what you will be needing, most electronic music systems are "modular" or flexible in some other way, so you can add to and modify your system with ease. You can usually start with a standard setup, work with it, and add additional "modules" as you find them useful.

The second area of planning is the acquisition of the parts and supplies you will need. This is not difficult, but it does take some time, and a missing item can easily hold up the whole works. Therefore, you will find considerable discussion of this area in what follows.

The third area of planning is to arrange an area to do your work in. The bad news is that unless you live alone, the kitchen table will not do, and just about any area you choose will become crowded within a short time. If you already have a good shop and workbench, that's great, but otherwise, you are going to have to make a special effort to arrange things so that you can work easily and have an area that you can use for an extended period of time. Actually, you will probably need three areas. First, there is the electronic construction area, the main area where you have parts, soldering, test equipment, etc. This is the main area, but you will also need an "etching" area and a "drilling" area. The etching area is the place you will etch your printed circuit boards. An out-of-the-way bathroom is good, and keep the etchant solution tray inside a heavy cardboard box lined with old newspapers. You must be very careful with the etchant solution, not because the chemicals are particularly dangerous, but because they stain anything porous (wood, cloth, hands). The drilling area is probably your shop or garage (perhaps a driveway in good weather). The main consideration here is to have a place where metal filings may fall without making too much mess.

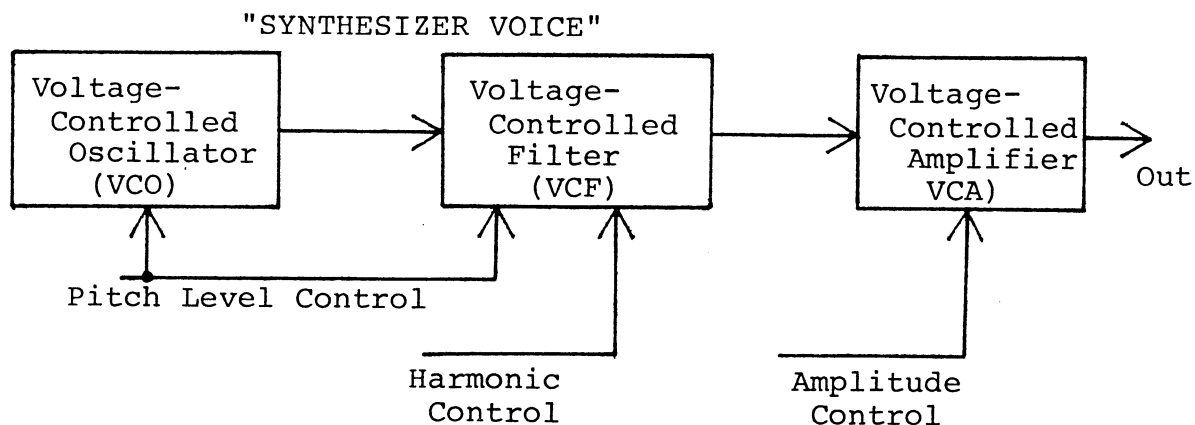
Next we will look at systems planning and parts acquisition in more detail.

## 1-4 DECIDING WHAT WILL BE IN YOUR SYSTEM:

In basic terms, this planning step is that of deciding how many VCO's, VCA's, VCF's, and so on will be included in your modular system. A VCO for example is a Voltage-Controlled Oscillator and is a module, one building block of your system. Modules are interconnected in what is called a "Patch" and this patch may be completely open (to be user determined with "patch cords") or it may be "pre-patched" or "hard wired" as a permanent setup. Only the individual builder/user can determine if he wants his system patchable or prepatched. The serious composer/experimenter will probably prefer a patchable, expandible, open system while a concert "rock" musician may prefer to have his standard sounds as prepatched features. The circuits are virtually identical in either case, so you don't have to change much in this area. The packaging problem however is different. For a patchable system, you will need a large "frame" to hold the panels as you build and add modules to your system. This frame is often found in a studio setup, and is probably large, sturdy, and heavy - not the type of thing to carry around. The prepatched synthesizer on the other hand is smaller, has only one panel, and is virtually a "plug in and use" unit. It is difficult to add things or make changes, so you have to determine your features before you build in this case. If in doubt, you can probably copy the features of your favorite commercial synthesizer, and perhaps leave room for an extra module or two.

A typical electronic music module has one or more inputs, one or more outputs, some panel control knobs, and perhaps a panel switch. When we consider a system of several modules, and the fact that all inputs are standard, as are all outputs, it is clear that possible patches become very numerous very rapidly. This is in fact, the whole idea of a patchable system - there are many many things to try. Yet at the same time there are many things that don't work all that well, so it is necessary to have certain "standard patches" to serve as starting points. The main standard patch that we will discuss is what has become known as the "synthesizer voice" where voice is used in the musical sense - as for example the voice or musical line provided by a trumpet or a violin. Perhaps more simply, it is a set of modules capable of being used as a solo.

The standard patch or synthesizer voice, which incidently is usually the main scheme of the prepatched units, is shown in the figure below:



The synthesizer voice consists of three modules, a voltage-controlled oscillator (VCO), a voltage-controlled filter (VCF) and a voltage-controlled amplifier (VCA). There are also three control voltage lines associated with the voice, and these are supplied by some sort of "controller" which is typically a keyboard with associated envelope generators, but may also be any of several other devices, including a computer, suitably interfaced.

The VCO is the source of excitation for the voice. The pitch of the VCO is controlled by the pitch level control voltage, and the output is a suitable waveform, typically of high harmonic content. The second module in the string is the VCF which processes the waveform from the VCO. It is controlled by two control voltages. one of which is the pitch level control, which serves to keep the VCF on the same pitch level relative to the VCO. Yet, if there were only this one control, the VCF would merely filter the VCO waveform in a static manner. Thus, in order to obtain a "dynamic" variation in the harmonic content of the waveform, it is necessary to have a second control, one which varies during any one "note". This dynamic control, relative to the level of the VCO pitch, is the most important key to the success of the standard voltage-controlled synthesizer. The final module in the string is the VCA. It serves to control the amplitude level of the VCF output. It is both a means of turning the note on and off, and of controlling the amplitude in a gradual manner, just as natural notes from acoustic instruments will not suddenly appear and disappear, but rather build up and fade down.

First we will begin with a description of the basic types of modules, and say a few words about what they are for. Then, for the actual selection, you can go to the reprint given in Appendix A.

### POWER SUPPLIES

Power Supplies are not really modules, but rather units that just feed all other modules their basic needs. They are the "energy source" of the whole system. Actually, the energy comes from the wall socket as usual, but an electronic music system, like most pieces of electronic equipment, do not use the wall voltage directly. Instead, a power supply converts the wall voltage to the desired levels. The most usual supply provides +15 volts, and -15 volts. You have to have one before you can do anything else.

### CONTROLLER INTERFACES

You want to control your equipment. Most likely, your controlling device is a keyboard, or some other mechanical device. A controller interface is just a device to make the mechanical unit speak electronics. For example, for a keyboard, the interface provides a voltage that indicates which key is down, when the key is down, and when a key is changing. Your exact needs will depend on your control device.

### VOLTAGE-CONTROLLED OSCILLATORS (VCO'S)

VCO's may be the most important part of your system, or at least a very fundamental part. The VCO is a device which transforms a voltage level into a waveform of a predetermined pitch. Look at it this way. You may be familiar with a manually "knob" controlled oscillator. You change the frequency or pitch with a twist of the knob. This is no way to make discrete jumps (as you get by pressing different keys on a keyboard), or to control the oscillator faster, or with more precision than you can do by hand. With the VCO, the job of changing frequency is transformed to a job of deriving the necessary control voltage.

### MANUALLY CONTROLLED LOW-FREQUENCY OSCILLATORS (LFO'S)

There is absolutely no reason to have a manually controlled oscillator. A VCO with its normal initial frequency knobs does the job just fine. However, VCO's are much more expensive and more difficult to build than LFO's, and a LFO or control function is often used. Hence, a LFO often does make sense. Often they can be used to fill up some unused space.

## VOLTAGE-CONTROLLED AMPLIFIERS (VCA'S)

VCA's are amplitude of "loudness" controls of the system. A signal input to the VCA is controlled by a control signal voltage, and the controlled version appears at the output. While their main function is amplitude control of the final signal, they may appear at other points in a patch to dynamically control some parameter of the sound. The VCA is actually an electronic multiplier circuit. It is generally a two-quadrant multiplier, meaning that the input signal is allowed to take on positive and negative values, but the control signal is permitted only positive values. The two-quadrant multiplier has an advantage over the four-quadrant multiplier in that it does not require a precise zero for shutoff, any slightly negative value will do. It may also be cheaper and provide better off-rejection than four-quadrant units.

## ENVELOPE FOLLOWERS

Envelope followers are included here because they are the reverse of a VCA. However, in general, an envelope follower is something not found in most synthesizers. You may want one however. An envelope follower derives a control voltage corresponding to the amplitude or loudness of an input signal. It is useful as a control device. For example, you can input a microphone, shout or sing into it, and the envelope derived can be used to control a synthesized signal. This works in general without regard for the actual material input through the microphone. Note that this is a loudness follower, not a pitch follower. Pitch followers are a bit too complicated to be included in this guide at the moment. Interested readers will find numerous suggestions in Electronotes.

## VOLTAGE-CONTROLLED FILTERS (VCF'S)

VCF's are very important (in fact essential) to the electronic music synthesizer as we know it today. A VCF is a filter which alters an input waveform, changing its harmonic content, and this alteration is in turn altered by a controlling voltage. You can make a system without a VCF, and you can produce sounds of the correct pitch and loudness, but the timbre or tone-color of the sound will be static. Traditional acoustic instruments tend to have changing tone color throughout a given note, and this is the function of the VCF. VCF's provide the "woooooow" sound of music synthesizers.

## TIMBRE ENHANCEMENT DEVICES (TIMBRE MODULATORS)

Timbre modulators are devices which change the harmonic content of a waveform, just as a VCF does. A VCF is a timbre modulator, but it is so important and such a large class that we keep them separate both in our mind and in our literature. A timbre modulator is a device which modulates timbre and which is not a VCF. Timbre modulators can be very useful, and often perform functions as dynamic as the function of a VCF, and they can have important advantages economically (they may not require expensive parts for tracking).

## ENVELOPE GENERATORS (EG'S)

EG's, sometimes called contour generators or transient generators, are very important in most synthesizers. They supply voltages that vary in time according to a predetermined pattern, after being initiated by a single electronic event such as a "gate" or a "trigger". Thus they are sort of memory units, and we can recall and use these envelopes whenever we want. For example, an amplitude envelope might be required that gives a note a duration of 200 milliseconds (0.2 seconds) and which allows the amplitude to rise and fall gradually. To prepare this, we would set the proper rise, fall, and duration times using the knobs of the envelope generator module, and then, we get the envelope out (to control a VCA) whenever a gate or trigger appears. Typically a gate or trigger is an automatic signal from the keyboard that appears whenever a key is pressed. Once set up, we get the proper envelope just by pressing a key.

## BALANCED MODULATORS (RING MODULATORS)

A balanced modulator or "ring" modulator is, like a VCA, an electronic multiplier, a four-quadrant multiplier in this case. It is useful in that it has two inputs which are identical (neither is a signal or a control per se) and one output. The output is the product of the two signals, and may be very complex in its harmonic structure. In fact, the ability to produce non-harmonic sounds is one of its most important features. Many think of it as a "special effects" device.

## FREQUENCY SHIFTERS

Although a few commercial frequency shifters are available, they

are not common synthesizer modules. Typically they will cost five to 10 times the cost of a standard module, and have a correspondingly larger construction time requirement. They are devices which take each and every component frequency in a waveform and shift it up or down by the same number of Hertz. They are very effective for voice and other live signals. Like balanced modulators, they may be classed as "special effects" devices, but they are also very useful for certain patches using tape delays and feedback.

#### SAMPLE-AND-HOLD UNITS

S&H Units do just what their name implies - they sample a voltage and hold it. Periodically given sample commands, they will break any continuously varying waveform into discrete steps of a constant level. These can then be used as control signals. A common application is to the sampling of a noise source. Noise is generally rapidly varying (so-called static noise) and seems smooth and uniform. Yet it is actually composed of fine structure which the ear smooths out. If we apply a S&H unit, we capture a small detail of the instantaneous fine structure. If the S&H drives a VCO for example, you get a series of random (but constant over their duration) pitches.

#### SLEWING CIRCUITS

In some ways, slewing circuits do the opposite of what a S&H unit does. Common slewing circuits are "portamento" units and "lag processors." A slewing circuit will move only just so fast. If you feed it a discrete jump in voltage, it will start in the required direction, but move slowly, giving a smooth variation. Thus, a sort of glide can be achieved. Generally a portamento is thought of as associated with a keyboard, while a lag processor is more of a general purpose module.

#### NOISE SOURCES

A Noise Source is a simple no-input, one-output device that supplies noise as a basic raw material for our synthesis purposes. If you listen to the output, it sounds just like the static noise you hear when an FM radio is tuned off station. This may not seem like it would be worth having, but it is often used for the synthesis of percussive sounds, whistling sounds, and for a variety of special effects. Also, as with the discussion of the S&H units, the noise can be slowed down and used as a source of random control signals.



## ANALOG DELAY LINES

Analog delay lines are relatively new devices which are a simple time delay, much as can be achieved with multiple playback heads on a tape recorder. They find their way into commercial "Phasor" or "Flanger" units, and often are used for special effects. Probably they would be widely used for choral effects generators were it not for their relatively high cost. Most builders would likely use an ADL unit in a phasor or flanger "comb filter" arrangement. The effect is also known as "Jetsounds" or a "swish."

## ANIMATORS

Animators are devices which attack a certain basic problem with electronically generated sounds: they tend to be too regular in the steady state. An animator introduces variations in the waveform so that it is already more interesting even before all the processing is complete. Most of these devices are new, and their applications are just now being explored.

\* \* \* \* \*

When it comes to choosing a system, you should do some careful study. However, keep in mind that very few well constructed modules will ever turn out to be a complete waste. As the number of modules goes up, the number of possible patches goes up even more rapidly. Thus there is seldom a need to worry that you won't be able to use something - just that you may wish you had something else at a given moment.

For more information on selecting system modules, see the reprint in Appendix A.

## HOW TO OBTAIN THE PARTS YOU NEED:

Obtaining the parts you need is not all that difficult (assuming you have the funds), but it does take a little effort, looking through catalogs and making out orders, etc. The first step, which you should take immediately, is to obtain the necessary catalogs. First, go out and get a handful of post cards, and also pick up one or more of the latest issues of Popular Electronics, Radio-Electronics, or Byte. The magazines you will use to locate dealers of mail order parts. The post cards you will use to request full catalogs from those dealers who do not give enough information in their ads in the magazines. Also, check out the list of suggested dealers, which is attached, and is typed so that you can photocopy it, cut out the addresses, and tape them on your post cards. We also sell a few parts that are used in our projects which are hard to get elsewhere. Subscribers to Electronotes receive our order forms from time to time, and receive lower prices and priority on the parts we have available. Non-subscribers may write for order forms and details.

There is a useful strategy for obtaining parts, based on the premise that the most important thing is to have the parts you need at the time you need them. Since most electronic music systems are modular to some degree, it is often possible to work on another part of the system if you don't have the parts you need, so in this case, things may be a little easier. In any case, you should follow three rules: First, don't order all your parts at one time from the same dealer. Secondly, price is important, but very likely there is not much difference between dealers, so there is less reason to consider price. Finally, be organized in your effort.

On the first of these rules, it is probably impossible to order all your parts from one dealer anyway since no dealer will have all you need. Secondly, no one dealer is likely to be able to supply a complete order on a wide selection of parts - some will be back ordered. Thirdly, you won't be able to determine ahead of time exactly what parts you need, because your plans are likely to change, and you may burn out a part here and there. As a rough guide, first separate out the more expensive and harder to obtain parts, and order these first. These are the parts

which are used for specific circuits (unlike op-amps for example which appear almost everywhere), which are so expensive you really can't justify spares for, and which you can not redesign the circuit around if you don't have them. Other parts, such as op-amps are so common that you will need a great many of them. You do not need to order all at once (unless you are ordering an excess for stocking purposes). Add enough of these commonly used parts to the orders you have already started so that you can get going. After working up these orders for the major parts and enough minor parts to get going, you will probably have orders to about four or five dealers, and will have arranged for about half to 3/4 of your parts. This is enough initially. You can make additional orders as you find parts you overlooked, as you need additional common parts, and as your initial orders come in with "holes." Expect at least five more orders before you ~~are~~ finished. As we said in the second rule, don't worry too much about price. A glance at two or three catalogs will tell you the going price and you will be able to spot a price that is way too high.

We also say as the third rule that you should be organized in your ordering. To us this means having your own order forms. Why? Not because you want the dealer to be impressed into thinking that you are some big company, but because you want him to be impressed with your neatness. He will be so impressed he will likely pull your order out of the pile to handle first (leaving all the messy ones until his mood improves), and he will do a better job of getting it right. Of course, if the dealer has his own form, use it, but you can still use your forms for organization and record keeping. But do make up the forms because in many cases you will be ordering out of magazines, and there will be no order form there. You may wish to use the form we provide (Appendix B) and you can remove it, type your own name and address on top, and make as many copies of this as you need. Note that the form has places for tax and shipping. Don't neglect these. Remember that the dealer has no choice about tax, so don't begrudge him that - if he didn't have too, he wouldn't. Shipping is another matter, but do it as he says. If he wants 15% added for shipping and handling, give it to him, but do begrudge him that. On a \$100 order, this would be \$15, totally ridiculous for shipping of course. It's really a 15% surcharge on all parts isn't it? Keep this in mind when comparing prices. On the other hand, some dealers give generous discounts on large orders, so keep this in mind when comparing prices too.

It should be realized that in general, the cost of semiconductors is going down. You should have little reason to stock up very far ahead. The same cannot be said for other items such as controls, switches, and particularly, packaging. Things like cabinets and metal panels are just expensive. This is also true of things like knobs and plugs which you will not always think of first. You will have to be prepared to spend as much on packaging and "cosmetics" as you do on the actual electrical components. Or, you can take the cheap way out!

The cheap way is to give up not only on a professional internal appearance, but on an external one too. You can make cabinets of wood and recover other items from electronic scrap. Never overlook a (carefully selected) garbage can as a source of the things you need. The throw-away piles from industrial and research buildings can yield a wealth of useful junk. Many such places throw away circuit board scraps, wire cuttings, panels and chassis units (often virtually undrilled), even transformers and the like. Don't plan to keep all the stuff you salvage, but even a useless piece of tube gear can be torn apart for the nuts and bolts.

Successful and economic parts acquisition is a matter of care, hard work, some experience, and at least some luck. Get started at it.

Some additional aids are found in Appendix B.

## **1-6** BUILDING YOUR SYSTEM

We look at the building phase of your project as divided into the following phases:

- (1) Preparation of Circuit Boards (35%)
- (2) Mounting of Parts on Circuit Board (30%)
- (3) Preparation of Packaging (15%)
- (4) Mounting of Circuit Board in Package (10%)
- (5) Testing of Unit (10%)

The percentages following the phases will give you a rough idea as to the total effort involved in this stage. Of course, some of these may vary with your individual style. In particular, if you are doing a super professional looking job on the panels, step 3 is going to take a lot more than 15%.

One of the nice things about this multi-phase view is that you can jump from one to another by virtue of the fact that you are building a modular system. In fact, if you are just beginning, it is suggested that you do take one module all the way through first so that you will discover and solve any possible problems before you are stalled by having all modules at the same state of construction when a problem comes up.

There has been a lot published in Electronotes and our related publications that concerns the building phase, so we will offer these earlier articles as reprints in appendices. We will now look at each phase individually.

### **1 PREPARATION OF CIRCUIT BOARDS**

We strongly suggest that you do not look for others to make your boards for you. Some commercial versions of boards for Electronotes circuits have been offered from time to time, but for the most part, you will have to rely on your own resources. We don't think this is much of a problem. Part 2 here will describe a simple method that we hope you will decide to use.

### **2 MOUNTING OF PARTS ON CIRCUIT BOARDS**

This part of the job, described in Part 2-3 and also in Part 2-7 is probably one of the most fun. It is the one where you will feel you are getting the most done. About all we can add to what we have

said in these write ups is that it is a good idea, where practical, to test the circuit as you build it. For certain, there are circuits which you will not be able to begin testing until the last wire is connected to the last switch, but consider the following: Even if you have to go to a little trouble, running power supply lines, putting in a fixed resistor where a pot will eventually be connected, and so on, it is often a lot easier to make a repair while the board is free on your bench than it is when it is mounted. Also, the psychological advantage of knowing that at least something is all set and ready to go should not be underestimated. So, test as you go if you can.

### 3 PREPARATION OF PACKAGING

In short, this step is that of finding a "home" for your boards. We discuss this matter in Part 2-4. You should feel free to do a neater job if you feel it is necessary. The only thing we have to add is that you should think big if possible. If you have the choice of making a setup bigger, do it. You will be surprised how big a system may become after a few years.

### 4 MOUNTING OF CIRCUIT BOARDS IN PACKAGE

Now that you have your boards made, and have prepared their package, it should be a simple matter of putting them in. It is. However, do not underestimate the amount of time this may take. If at all possible, wire as much of the panel as possible before you put the board in, and keep panel-board wires to a minimum. It's neater this way. You might suppose that you have no real choice in this matter (there must be interconnections!) but there is a tendency to go back to the board more often than is necessary. If a wire comes to the panel in one place, and you need it again on the panel, wire across panel, not back to the board.

### 5 TESTING OF THE UNIT

Oh, if it were only possible to say something rational and sure about this! Connect it up and hope. If you have done things right, the odds are it will work. However, the odds are also that you have probably missed something, and the odds are virtually certain that in a system consisting of say 10 or more modules, you are going to find one that is not going to work the first time. In short, you are going to have to know how to troubleshoot. Common sense will get you a long way here, but if you need help, read over the notes in Appendix C.

## THE ORDER OF BUILDING THINGS

We have said little about which modules to build first. Does it make a difference? Not really, but if I were beginning, I would want to get as many results as soon as possible, even realizing that I was not going to get the full show until I get quite a few modules done. Thus I would start with a VCO because I could listen to it and its various waveforms. Then I would build either an LFO to modulate it, or a VCF to filter it. If possible, always try to have something new and interesting to try as a reward for finishing a module.

## SYSTEM INTEGRATION

System integration will be the most enjoyable part of your whole construction effort. It is concerned with putting all your units together, neatening things up, making sure everything is working, and then just playing around a bit. It also involves a certain amount of just sitting back and admiring your own handiwork.

Specific things to look at during system integration are:

- (1) Are the modules arranged so that the most common patches will be patched with the shortest cables?
- (2) Are all jacks and controls properly labeled?
- (3) Are the ranges of inputs and outputs properly set so that initial controls can be easily adjusted to the desired operating ranges?
- (4) Is the power supply driving all the necessary modules without overheating?
- (5) Are patch cords of the proper lengths and in good shape?
- (6) Is the system documentation in good shape so that it will be useful when needed in the future?
- (7) Are there any shortcuts or temporary setups that you must remember to improve before too long?
- (8) Are the interconnections to your audio equipment well established and convenient?
- (9) Is the "studio" area set up so that you will be able to work with and experiment with future additions?



## STANDARDS USED

The following standards are used in most of the designs and schematics that follow:

Power Supplies: +15 []; -15 []; Ground [], +5 []

Signal Levels:  $\pm 5$  volts

Envelope Levels: 0 to +5 volts

Gates: +5 = ON, 0 = OFF

Triggers: Short +5 volt pulses

Voltage Control of Pitch Level: 1 volt per octave into 100k input impedance

Input Impedances: Greater than or equal to 20k (100k or greater preferred)

Output Impedances: Output from op-amp through 1000 ohm as standard (or equivalent), except on 1v/oct. control lines which are direct from op-amp.

Multiplying Modules: Unity gain at 5 volt levels

Notation: Standard schematic symbols for the most part.

Connections shown as:  or  or 

Jumps shown as: 

Cross is avoided except where showing jumps would be too cluttered

In many schematics: PC = Panel Control  
TP = Trim Pot

Power supply connections to op-amps are generally +15 and -15, but are almost always not shown. Where they are shown, it is usually because a non-standard setup is used.

## SUBSTITUTION OF OP-AMPS

In the circuits that follow in this manual, you will find many that specify certain op-amps by type number, and some which do not specify any at all (just the op-amp symbol). There are two things to be remembered here. First, to a first approximation, all op-amps are the same. Secondly, however, at times it does make a difference which op-amp is used.

To understand op-amp substitution, it is necessary to understand why a designer has chosen the one he did. It may be he choose it because he had it on hand. It may also be that he needed one with a certain characteristic (such as high speed, or low bias current), and chose one of several possible devices somewhat arbitrarily. In still other cases, the choice is somewhat arbitrary, but he may have a subtle reason for choosing one type over another (often based on personal prejudices and past experience). In any case, it is possible, often desirable, and at times almost totally necessary to make op-amp substitutions for the types indicated.

To give you a better idea, it will be useful to review the history of the IC op-amp ever so briefly. Back about 1972, we had mainly the type 741 to use. We thought they were great then. Now we think of them as slow and as requiring a very large input bias current! Yet they are still excellent choices in some designs where nothing better is required (In some cases, using a better op-amp is possibly asking for trouble; an op-amp that is slow is not going to give high level high frequency oscillations, no matter how hard you try to make it.) A bit later, the type 307 became available to us. It was lower on input bias current requirements, but no faster than the 741. For the lowest bias requirements, we used either a FET input device (such as the 536) or an external source follower. Next along came the MC1456/MC1556 which was typed as 556. This was five times faster than the 741/307 types, and also improved further on input bias current. It still meets our non-FET needs, but costs about a dollar. Next we had the CA3130, a CMOS (FET-input) type, which would run only on a max supply of 15 volts. The CA3140 works on full  $\pm 15$ , but the output stage has a minor defect for our purposes, and was only used internally in modules. Finally, along came the BiFET types which are fast, low bias, and cheap (about 50¢). We have generally chosen the LF351 as our BiFET op-amp, although many others are available, and can be used.

In looking through these circuits, the reader will perhaps understand how a certain op-amp was used because it was the best available at the time. What about today? As you might expect, we would probably use a different op-amp in many places today. We could go through and change the numbers, but this would attack only part of the substitution problem. For example, if we specify an LF351, it could be for speed or bias current (or both). If it is only for bias current, you could well use a slower (and cheaper) BiFET type like the LF13741. For this reason, and as a side benefit, to help you understand the design, we will be adding "dot codes" to many circuits. These dot codes will add a minimum of additional clutter to the diagrams. They will consist of two dots (or open circles) side by side, with the first dot representing bias current, and the second dot representing speed. If the op-amp was selected for low bias current, the circle on the left will be filled in, and/or if it was selected for high speed, the circle on the right will be filled in. Remember that the sequence is bias-speed, or b-s left to right. Below we show the possible combinations, and suggested op-amp types:

○○ No particular requirements. Use 741, 307, or types below

●○ Needs low-input bias current. Use any FET-input type. The LF13741 (commonly available as house number SL32005) is an excellent choice if available for much less than the LF351 or other BiFET. CA3140 is also acceptable if the output of this op-amp does not go to an output jack.

○○ Needs high slew rate (high speed). For signal levels, types like 741, 307, and the LF13741, are just marginal in speed for  $\pm 5$  volt signal levels. For more margin, or for higher signal levels, use the LF351 or other similar BiFET (slewing at 2v/microsecond or faster). Use 556 if you have them left over.

●● Needs high slew rate and low bias current. The LF351 or similar BiFET is a good choice.

NOTES: Certain types such as 748 and 301 are used because they can be uncompensated and used as comparators. 308's may also be special purpose devices relying on a certain compensation. If you prefer, you can choose packages that contain more than one op-amp if they are similar to the above. Devices without dot codes are just ones that don't have dot codes, for one reason or another. It may be that you should not substitute, but you must use your own judgement. See Electronotes application note AN-100 for more op-amp type listings and groupings.



The following 10 pages are a reprint:

"Modular Design - General Considerations, Inputs and Outputs"

-from Chapter 5a, Musical Engineer's Handbook

CONTENTS:

Introduction

Voltage Input Structures

Timing Signal Input Structures

Output Structures

INTRODUCTION:

An electronic music module is a self-contained (except for power supply) functional module for use in an electronic music synthesis scheme. Thus, a module is a part of a larger synthesis system. It can be modular in the sense that it can be pulled out of one physical position and plugged into another, or the modularity may be more conceptual in the "block diagram" sense. What we have in mind here is a unit which has a physical position (not necessarily movable) and which has accessible electrical terminals. These terminals are connected to the terminals of other modules through the process of "patching." By using a modular system as opposed to a "hard wired" system, the user is able to realize a wider variety of synthesis schemes.

A module consists of two parts: the electrical functional block (the "guts" of the module) and its interface with the rest of the world. Often this simply comes down to a circuit board on the one hand and the front panel fixtures on the other hand. This is the sort of picture we have in mind when we discuss a general module.

The module's functional section ("submodule" or circuit board) is the most difficult part of the design, and each different type requires a separate discussion. This is the purpose of the remaining chapters in Section 5 and Section 6. In this chapter, we want to discuss the interfacing procedures, as these are common to most types of modules.

The main interfacing mechanisms are: signal inputs, control inputs, timing inputs, panel controls, and outputs. These mechanisms in turn can be grouped according to the type of electrical structure and signal routing that is used:

1. Voltage Input Structures: These voltage inputs may be either control inputs or signal inputs; the only difference between signals and control is the use, not the input structure. Generally, an op-amp summing node is the voltage input structure. In addition, often times voltages from the panel controls enter the functional blocks through the voltage input structure. For example, the voltage input for the control voltage of a VCO often serves to handle externally applied voltages, and "Coarse" and "Fine" tuning voltages from panel pots.
2. Timing Signal Inputs: For timing signals, often the voltage level is of no importance. The important thing is that the voltage meets certain conditions, and that the time at which it meets the conditions is well defined. For example, we might want a "Gate" to appear when the signal voltage crosses the 0.5 volt level. This input structure generally requires a comparator.
3. Outputs: Outputting a voltage is largely a matter of making some voltage in the module available to the outside without disturbing the inside. This is of course, a buffer. In addition, it is often useful to have an output with certain "standard" properties, e.g., a standard 1k output impedance.
4. Individual Panel Controls: These are controls that are connected to the insides of the functional block directly, not through another input. An example might be a High/Low range switch.

We are thus left to consider shortly the details of the input structures and output structures. We shall see that there are logical and standard ways of doing these, but the actual combination of structures required in a given module will require individual attention. This is no small point, because the interface between modules is also the interface between the electronics and the musician, at least as far as the general type of synthesis process is concerned. Setting up this interface properly is one of the two main ways (the other being the controller) by which a musician makes the electronics make music. It is thus the interface that "shows" and this must be carefully considered. The main considerations are:

1. The exact combination of structures required: For example, a VCF needs a signal input, at least two control inputs, and one or more outputs as a minimum. A noise source on the other hand has no inputs at all (by its very nature) and the only interface structure other than an output or two would be some sort of control that controls the "character" of the noise. This requirement thus has a natural (what can be done), and a use (what does the specific user want to do in this specific case) aspect.
2. The panel layout: This involves the questions: What is a logical panel layout from the designers point of view? What is a logical layout from the musicians point of view? When these two layouts differ, is the musician better served by being forced to use the designer's choice?

### 3. The Patching Method: Plugs and jacks? Matrix switch? What?

These are all things that are important to musicians. Often the musician will not be in a position to appreciate the great difficulty you went to to develop a stable VCO. He will of course "appreciate" the "value" of a VCO that drifts badly. Suppose however that you do give him a good one; how is he to know that it is even possible to make a VCO that does drift? With the interface however, the musician will appreciate a set of controls and jacks that are conveniently organized, and which allow him to set up a desired patch without elaborate contrivances.

For the exact combination of interface structures required, it is suggested that the individual unit be considered carefully to see if all useful manipulations have been implemented. Secondly, all standard or expected patching systems should be considered. Thirdly, more structures than this minimum can be added as cost and panel space will allow.

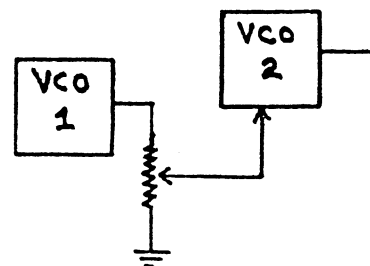
The question of panel layout is one of utmost importance. There are seldom any critical circuit layouts that dictate a certain panel layout, so the designer is in general free to choose the most logical setup. It is suggested that the designer choose a panel layout according to what he knows about the logical organization of the functional block. If properly done, this layout should be so logical that it would not be necessary for the designer to even label the jacks and knobs [of course he will do so for others!]. Using what seems like a logical layout to the designer who designed the actual "guts" of the thing has the advantage that the actual device and its capabilities will be more naturally represented, and this will be passed on to the user. Nonetheless, panel layouts from other synthesizer modules should be studied, as there may be some sort of standard that should be used, or the designer may find that someone else's conception of the logic is better.

The actual patching system will depend a lot on the eventual use. For most purposes, standard 1/4" phone plugs and jacks are recommended. Many users of this book will be working on the expansion of existing systems, and these plugs and jacks are the most common for interconnections. Aside from that, plugs and jacks are the most conceptually simple patching system, and the most reliable system. Where panel space must be held to a minimum, smaller jacks or matrix switches can be used.

### VOLTAGE INPUT STRUCTURES

The voltage inputting structure that we want to employ is the op-amp summing node. This structure is useful for both control inputs and signal inputs. The standard inputting resistor is 100k ohms.

What we are actually considering is the transfer of one voltage to the input of a module. In many cases, we want this transfer to be incomplete - that is, we want to attenuate the voltage, and use only part of the available input. Thus, we must make some provision for attenuating the signal somewhere in the path. First of all, we want to make the case for the use of input attenuation. The device we will use to attenuate a voltage is almost always a variable resistor (a potentiometer). In the system shown at the right, assume that the output of VCO-1 is a  $\pm 5$  volt sine wave (a peak-to-peak voltage of 10 volts). If we connect this to a standard VCO control input (to VCO-2), this would produce a swing of 10 octaves. Suppose we want only a small degree of frequency modulation. We need to reduce the signal from a 5 volt amplitude to a smaller value (e.g., 1/12

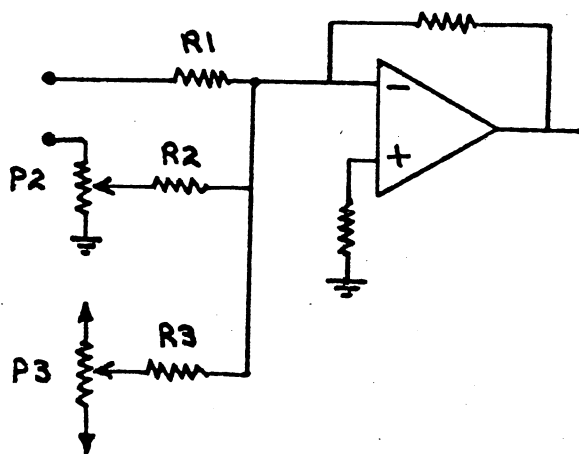


of a volt for a swing of  $\pm 1$  semitone). We can simply insert a pot in the transfer line to form an adjustable voltage divider.

It is clear that this pot is a useful control, and we want to make it a permanent part of the system. The question then comes up as to where the pot should go. Should it be a part of the output of VCO-1, or part of the input of VCO-2. Well, we shall see that it should be part of the input structure of the driven module, not part of the output of the driving module. The basic reason for this is that the voltage level should be adjusted at the point of use. Suppose you want to use the output of VCO-1 to control two different VCO's. You want one to have a semitone swing of FM, and the second to have 3 semitones of swing. Once you have cut the voltage to  $1/12$  volt for the first, there is no way you are going to get the  $3/12$  volt for the second. Thus, the input attenuator allows each use to select whatever part of the maximum signal is required for the individual application, and still allows the voltage to be available at full value for other applications.

The summing node input structure has been studied in the chapter on the use of op-amps. The important consideration here is that the summing node is the - input of an op-amp that has negative feedback. The summing node remains at ground potential as long as the + input is grounded. Any excess current or current deficit into or out of the summing node is made up by the current through the feedback resistor. The overall device is an inverting, weighted summer.

We can now consider a typical input circuit as shown at the right. The input connected to  $R_1$  is a standard 1 volt/octave control input.<sup>1</sup> Generally,  $R_1$  is 100k. This input would handle the main control voltage from the main controller of the system. For proper matching of such inputs as used on VCO's and VCF's, this type of resistor should be a 0.1% tolerance type - if there is more than one of this type on the module. Note that this input is used without attenuation. The input supplied through  $R_2$  does have an attenuator on it. If the resistor  $R_2$  is 100k, then the attenuator adjusts the response from 1 volt/octave down to zero. This is how microtonal scales could be generated. This type of input



is also useful for inputting envelopes or any general control signal. This is the type we had in mind for input attenuation. Note that while  $R_2$  could be 100k, it can be any value in a much wider range. For example, if it were 33k, the response would range from 3 volts/octave down. There would be less resolution, but a greater range. It should be realized however that these inputs have nothing to do with the overall range of the actual functional block, except for the fact that they may reduce it. The important thing as far as the functional block is the sum of the control voltages at the output of the op-amp. The functional block inside responds to this voltage. If the functional block only responds to a control sum from say -3 volts to +5 volts, then the range is limited to 8 octaves. If the control sum can only swing from -1 volts to +10 for example, then the range is cut to 6 octaves, and the control sum above +5 is wasted. The best way to determine the full value of the control sum is to connect up only one input to the module through a 100k resistor, and vary the input voltage from +15 to -15 and find out what the useful range is. The designer can then use the op-amp summation techniques to determine how to set the input resistors so as to keep the control sum in a useful range."



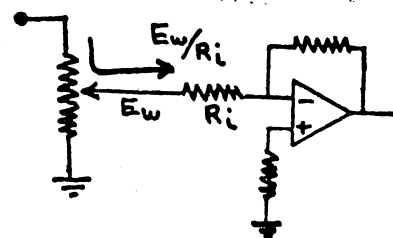
This leads to consideration of the third type of input resistor ( $R_3$ ) which is a range control. For a coarse control, the value of  $R_3$  should be selected so that it sweeps the full range of the control sum. It may be necessary to select a fixed resistor and connect it to either +15 or -15 to "center" this control. For a fine control, a value of  $R_3$  much greater than 100k should be used. To sum this up, the following values are suggested:

| Type of Input | Type of Control                  | Suggested Value    |
|---------------|----------------------------------|--------------------|
| $R_1$         | 1 volt/octave standard           | 100k               |
| $R_2$         | Auxiliary input (e.g. envelopes) | 33k to 100k        |
| $R_3$         | Coares                           | 100k or lower      |
| $R_3$         | Fine                             | 1.5 meg to 4.7 meg |
| $R_3$         | Ultra Fine                       | 15 meg to 22 meg   |

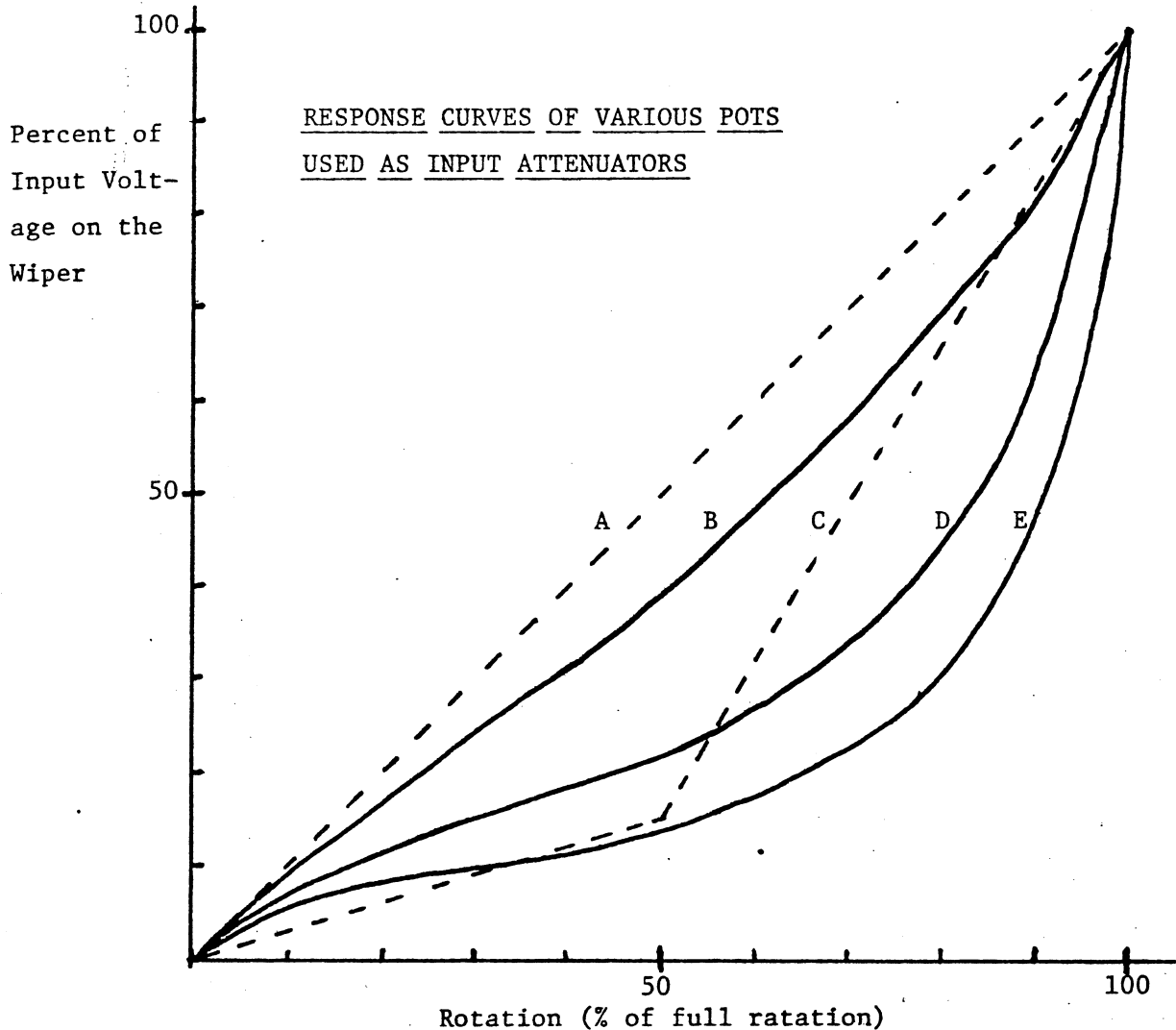
What we have said above was mainly directed at control inputs, but the summing node is also useful for signal inputs. Here, the  $R_2$  type of input is generally used with  $R_2 = 100k$ , and set so that with the pot set to the top, the standard signal level enters the circuit with unity gain. The pot is thus strictly an attenuator. For the control inputs, for values of  $R_2$  less than 100k, the summing node amplifier actually provides some gain (up to a gain of 3 for  $R_2 = 33k$ ). This is useful for many control signals where the envelopes may be at a maximum level of 5 volts, and you may want more than 5 octaves of swing for a VCF for example.

It is now time to consider the pots. What should the value of the pots be? Should they be log or linear pots? First of all, consider that the pots are acting as voltage dividers in the type  $R_2$  and  $R_3$  input structures. Thus, it is natural to suppose that the pots should have a value much less than the attached  $R_2$  or  $R_3$  resistor. This is the standard consideration for a voltage divider. On the other hand, the voltage on the pot wiper has to go through the full range from one extreme to the other, so it is not necessarily required that the pots have low values. A log pot is one which has two linear sections. The sections are connected in the middle of the rotation. The log pot provides about 15% of its total resistance over the first half of the full rotation, and the remaining 85% on the second half. This type of pot is also called an audio pot. The reason for this latter name is that these are the types of controls that are used as "volume controls" and they are made that way since the amplitude response of the ear is logarithmic. Thus, there are certain applications where more resolution is required for small signals, and these are applications for log pots. None the less, the log pot still makes available all voltages throughout its range. Thus to a certain extent it can be said that the choice of pot is arbitrary. The considerations are the trade offs between input loading (for an  $R_2$  type) and current consumed (for an  $R_3$  type) on the one hand, and the degree of resolution and accuracy on the other hand.

We mentioned above that a pot is accurate as a voltage divider only when its value is much less than the resistor loading the wiper. The situation is illustrated at the right, where it is assumed that a 100k input resistance is attached to the wiper, and that the other side goes to a summing node (ground). The input structure is such that the voltage summed is always the wiper voltage, but in order to report the wiper voltage, some current must flow, and this steals some current from the voltage divider, and the



voltage on the wiper is different from what you would expect from the rotation of the pot shaft. A linear pot will not have a linear response. We show below calculated curves for five different cases. Curve A is the linear case, and would occur only where the wiper voltage is buffered, or where the pot is very small (10k or less). Curve B shows how the response changes when the pot is made 100k. Curve D shows the case for a 500k pot, and curve E shows a 1 meg pot. This warping of response is not as serious as one might expect - at least for the 100k case. Most all controls are set by ear, not by rotation position. The question is not one of accuracy, but of resolution. Note for example that a linear 500k pot has a curve that is more like a log pot (curve C). Curve C is the ideal log pot case. Higher resistance log pots would of couses show warping from curve C just as linear pots warp away from curve A. Note that curve E might be awkward in that it has a near linear region for very small rotation, a fairly flat region (10% to 50%), and then a very steep rise.



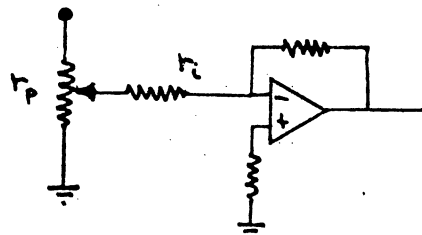
A.- Linear Curve: Applies to  $r_i \gg r_p$ , or for the case where the wiper voltage is buffered.

B.- Linear 100k pot [ $r_i=100k$ ,  $r_p=100k$ ]

C.- Log Pot for  $r_i \gg r_p$ .

D.- Linear 500k pot [ $r_i=100k$ ,  $r_p=500k$ ]

E.- Linear 1 Meg pot [ $r_i=100k$ ,  $r_p=1\text{ Meg}$ ]

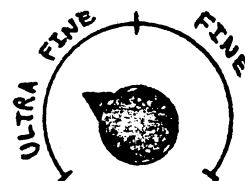


The following values and types of pots are suggested:

| <u>Type of Input</u>      | <u>Type of Pot</u> | <u>Suggested Value</u> <sup>†</sup> |
|---------------------------|--------------------|-------------------------------------|
| Auxiliary Input ( $P_2$ ) | Log*               | 100k                                |
| Coarse Control ( $P_3$ )  | Linear             | 100k or 50k                         |
| Fine Control ( $P_3$ )    | Linear or Log**    | 100k                                |
| Ultra Fine ( $P_3$ )      | Linear             | 1 Meg or 100k                       |
| Audio ( $P_2$ )           | Log                | 100k                                |

\* A log pot is selected for auxiliary control since resolution of small control voltages is often necessary, and many of the parameters that are controlled (amplitude and frequency) are subjectively logarithmic.

\*\* A log pot is often a good choice for a fine control. The reason for this comes when you consider that before you adjust the coarse control, you usually center the fine control, or else you won't have range on either side to fine tune. With a log pot, you can set this initial position in the center of either of the two linear regions as shown at the right. In this way, you can have two different degrees of fine tuning that differ by about 10:1 in response. With careful selection of the  $R_2$  type resistor to go with this, a very useful control is obtained.



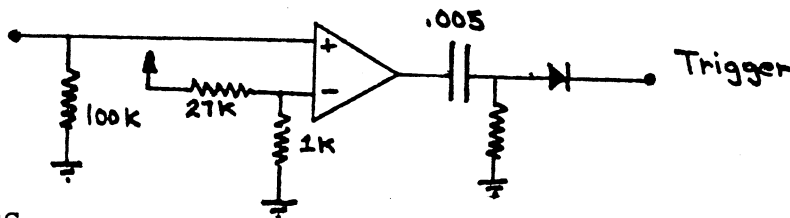
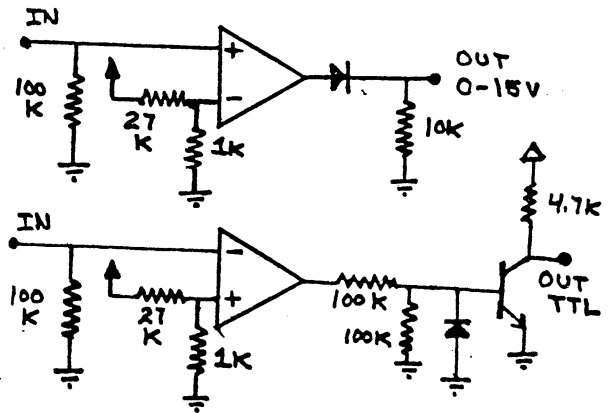
† Mostly 100k values are selected. Anything from 50k to 100k should be fine. While there is some warping of the rotation/wiper voltage curve with 100k, the main reason for preferring this value is that it causes only a 1% load on the standard output structures we will be using.

### TIMING SIGNAL INPUT STRUCTURES

Timing signals are such voltages as gates and triggers which tell modules when to start and stop their functions. The most common of these signals are the gate and trigger signals from a keyboard that normally control one or more envelope generators. It often seems like these could be just permanent behind-the-panel connections, but there are cases where it is very useful to break into these lines. For example, the keyboard trigger can be brought out, fed to the input of a sharply tuned filter and made to "ring" the filter. On the other end of the line, we often want to trigger an envelope generator from some source other than the keyboard. Thus, we consider the inputting and outputting of timing signals. The outputting of a timing signal is the same as any other voltage output and will be considered with the other outputs.

Inputting an "ordinary" voltage for use as a timing signal is a little trickier. The essential thing about a timing voltage is that it represents some event that is defined in time. We are thus left with the problem of representing a time point by a certain point on a waveform that may be quite different from the standard timing signals (sharp pulses and rectangular waveforms). The obvious device for translating a small transition into a sharply defined one is the comparator. For example, we can define a gate as occurring whenever the input voltage exceeds 0.5 volts. A trigger might be defined whenever a waveform crosses the 0.5 volt level from the negative side. For this purpose, the ordinary op-amp comparator or "weak" Schmitt trigger is usually sufficient. In the examples that follow, we will choose 0.5 volts

as a reference level, but other levels can be chosen. [Ground level is not chosen because too many signals have an off state very close to zero, and erratic triggering could occur.] The first example at the right shows a comparator used to define a gate whenever the input goes over 0.5 volts. The input impedance is 100k, and the output gate is either zero or about +13 volts. This is a useful gate level for CMOS circuits run between +15 and ground. If a TTL version of this is needed, it is possible to reverse the input terminals of the comparator, and employ the standard analog to TTL stage, which is inverting. This circuit is shown at the right. These same two circuits will work for inputting trigger signals if the pulse shape is established outside. In some cases, it is useful to obtain a trigger signal from a slowly varying input signal. In such a case, the comparator output can be differentiated to give a short pulse, and the unwanted transition can be blocked with a diode. The circuit below gives a positive trigger whenever the input crosses 0.5 volts from the negative side.

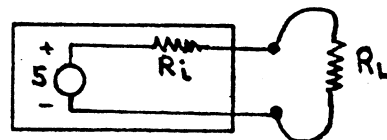


## OUTPUT STRUCTURES

The standard output structure for electronic music modules is a voltage output with a 1000 ohm output impedance. This structure starts with a low output impedance voltage source such as the output of an op-amp, FET source follower, or emitter follower, and 1000 ohms of series resistance is added.

Since it is the output voltage that is used, we must ask how the module being driven will change the open circuit properties of the module that is the source. We have seen above that the standard input structure has an input impedance on the order of 50k to 100k. Thus, we can see that the load can be expected to be a few percent of the output impedance.

First it is necessary to take a look at the meaning of output impedance. Suppose you have a black box with two output terminals and you measure the output voltage with a device that has a very high input impedance (scope or FET input voltmeter for example). You find the voltage is 5 volts, and you then connect a 5k resistor across the output terminals and expect a current of 5/5000 amps (1 ma) to flow. Actually, you will find a lower value of current flowing. This is because the voltage source has some impedance of its own. It is the usual practice to consider this to be a resistance  $R_i$  in series with an (Imaginary) perfect voltage source. This internal resistance can be very important in many practical problems. A familiar example is the ordinary flashlight battery which may measure a full 1.5 volts, and yet fail completely to light a bulb due to large increases of internal resistance as the cell ages. Suppose the black box we are considering is a perfect 5 volt source with a 1k series resistor  $R_i$  as shown at the right.



Two things become apparent when we look at the circuit as a series resistance problem and just apply Ohm's law. The current is actually 0.83 ma and the voltage on the output terminals drops to 4.17 volts, down 17% from the open circuit case. This is called loading. What happens when we consider other loads?

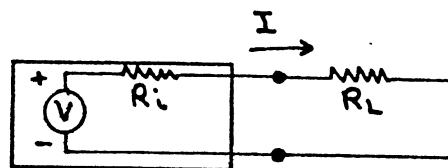
When we put a 100k resistor on the output terminals, we get a current of 0.0495 ma, for an output voltage that has dropped to 4.95 volts, or about a 1% drop. This is not too serious in most cases. Another interesting case is the load of 1k. In this case the voltage drops to 2.5 volts, exactly half the open circuit voltage. This is a common way of measuring output impedance - you just load down the circuit until the voltage drops to half its unloaded value. The load resistor and the output impedance then have the same value. It is nearly as easy, and perhaps easier on the circuitry, to just load it down by 10 or 20 percent, and use a standard series resistance calculation using Ohm's law.

As a variation, consider a very high output impedance (say 10 Megohms) and use the device to drive a series of loads from zero ohms to 100k. For zero ohms, the current that flows is 0.5 amps, and the voltage is zero (short circuit). For a load of 100k, the current is 0.495 amps, and the voltage is still very small compared to 5 volts (0.0495 volts). Thus, the current is very nearly constant for a wide range of loads, and the voltage is very low. This is the simplest example of a current source.

Thus, a low output impedance (relative to expected loads) is necessary for a voltage source. A high output impedance (again relative to expected loads) is necessary for a current source. The case in between, where the two are of similar magnitude should be avoided.

Since we are interested in inputting voltages for signals and controls, we have to be able to output voltages, and thus the voltage source (low output impedance) is used. Note that it is not necessary (nor is it desirable) to have input and output impedances matched. It is necessary to match impedances only when power is to be transferred, or for preventing reflections on transmission lines. We are transferring voltage, and we want to do it from a low output impedance to a high input impedance. This mismatch is desirable, but note that the other one (the current source) isn't. Transfer of power is the middle case. It is perhaps instructive here to show that the transfer of power is most efficient when impedances are matched. To do this, consider the circuit at the right. The power developed in the load is:

$$P_L = I^2 R_L = \frac{V^2}{(R_L + R_i)^2} R_L$$



This expression can be differentiated with respect to  $R_L$  to give:

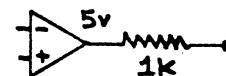
$$\frac{dP_L}{dR_L} = \frac{V^2(R_i^2 + 2R_i R_L + R_L^2) - V^2 R_L(2R_i + 2R_L)}{(R_i^2 + 2R_i R_L + R_L^2)^2}$$

Setting this equal to zero to get the minimum gives:  $R_i = R_L$ .

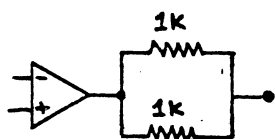
Getting back now to the low output impedance, it might seem like the very low (few ohms) output impedances of op-amps, emitter followers, or source followers would be the best we could do. However, there are several advantages to adding a 1000 ohm series resistance, and few drawbacks. For one thing, it protects

both the output and the inputs by limiting current in the event that something goes wrong. Op-amp outputs can usually stand at least momentary shorts, but it is not good practice to take advantage of this. It is virtually impossible to plug in any plug and not short the live contact to ground in the process. The 1k output resistor holds this in check. We shall see shortly that the standard output also allows some degree of mixing of signals. The errors that are caused by this internal resistance are as we have shown above, typically a few percent.

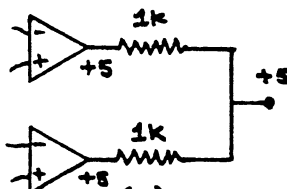
To show how output mixing can be achieved, consider the typical output stage (a) shown at the right. The output voltage is 5 volts, and the output impedance is 1k. If we now parallel the 1k output with a second 1k resistor as shown in (b) below, nothing changes, except the output impedance drops to 500 ohms. Nothing additional happens if we drive the second resistor with its own op-amp as in (c) below, as long as both voltages are the same. The mixing effect comes about when the voltages are different. A current now flows between the two op-amp outputs. Since the two resistors are the same value, and form a voltage divider, the voltage at the output is the algebraic average of the two op-amp outputs. Thus we have a simple 1:1 mixer (d).



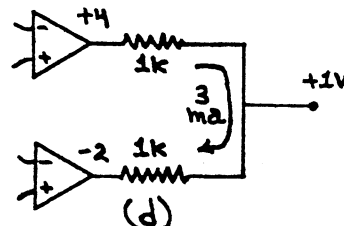
(a)



(b)



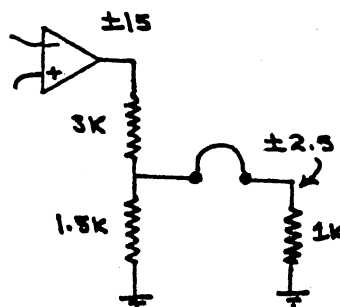
(c)



(d)

If three or more jacks are shorted, the average of all signals shorted is obtained. The main drawbacks of this method are: 1) The output impedance of the combination of  $n$  inputs drops to  $1k/n$ . 2) The mixing is always at equal levels: 50%-50%, 33%-33%-33%, etc, and you can't get something like 70%-30%.

Often, we have signals that are at voltages that are higher than the level at which we want to output them. For example, when waveshaping a smooth signal into a pulse, a comparator is usually used, and this output swings between  $\pm 15$  and we may want the output to be only  $\pm 5$ . To reduce this, we select the appropriate resistance ratio (2:1 in this case), and select actual resistances so that if the two resistors were placed in parallel, the resulting resistance would be 1000 ohms. For our example, we find that this requires that the resistors be 3k and 1.5k. The circuit is shown at the right. To demonstrate that the output impedance of this voltage divider is in fact 1000 ohms, we have shown it with a 1k load. The output voltage with a 1k load is 2.5 volts, which is half the open circuit voltage of 5 volts, and as we demonstrated above, this is the condition for the output impedance being equal to the load.



The 1k output resistors (or the equivalent voltage dividers) are generally used on all front panel output jacks. One exception to this is the main control voltage from the main controller of the system. This control voltage is standardized to be applied to a 100k summing node for a one-volt/octave response. Some sort of output jack for this main control voltage is usually necessary for interfacing with other systems even where the main distribution of this voltage to modules is by switches. If a 1k resistor were used here, it would cause the effective input impedance to the summing node to be 101k, a 1% error which would be very serious for pitch control.

## PART TWO

### CONSTRUCTION PRACTICES

- 2-1 INTRODUCTION
- 2-2 HINTS ON SETTING UP AN  
ELECTRONIC MUSIC WORKING AREA
- 2-3 HOW I MAKE MY P.C. BOARDS  
IN A MATTER OF A FEW HOURS
- 2-4 PUTTING YOUR CIRCUIT BOARDS INSIDE SOMETHING
- 2-5 MAKING RACK PANELS
- 2-6 SOME MORE SUGGESTIONS ON ACTUALLY  
GETTING CONSTRUCTION UNDERWAY
- 2-7 HOW TO ACTUALLY BUILD SOMETHING  
(App. Notes 14-18)





All of the material below is reprinted from Electronotes or from our Application Notes. There should be plenty of practical information on construction, and moreover, we hope to leave the impression that construction can be approached in an organized and rational way. As we have stated before, construction itself should be a major part of the fun, or else you should not even start a project. However, at best construction can get tedious and tiresome, and poor organization could make things downright frustrating, especially as it may lead to a unsuccessful effort. Take the time to read our suggestions, and then take the time to prepare, plan, and organize. It will be worth it.

## 2-2 HINTS ON SETTING UP AN ELECTRONIC MUSIC WORKING AREA:

-by Bernie Hutchins

We are discussing here what I will call an electronic music working area. Many would call it a workshop while others would call it a studio. It is in many cases more accurately called a playroom, because it is the most fun part of the musical engineer's work. The electronic music working area is the place where the equipment is actually set up. Properly set up, it is a joy to work there. Improperly set up, it may be difficult to work with and may even be dangerous. It deserves a good deal of consideration, and is often set up in a hurry and with little thought. Here we want to give a few hints which may be helpful.

The first rule I believe is that the setup should be flexible. No parts of the system should be physically (or psychologically) set in one place if this can be avoided. You may think you have the whole thing figured out, but then you try to do some actual work and right away find that the part you need to use is in an inconvenient place. My general rule would be that the first time you find something out of place, make a note to think about it. The second or third time the same thing happens, move it.

When it actually comes to setting up equipment, we find that things must be plugged into wall sockets or otherwise given power. The average electronic music system requires more sockets than are present in the average room. The result is often a conglomeration of extension cords and cube taps. This is certainly not very neat, and it may be dangerous. It is not dangerous in the sense that you may have an overload. Most electronic music equipment requires relatively little power, and even with a spider web of extensions, there is little current required. The danger is that the connections may become loose, and a dangling patch cord may drop on a half exposed prong. Certainly electronic music patches are not meant to handle raw 110 volt signals, and neither are people! As far as we know, there have been no serious (or at least no well known) accidents with electronic music equipment. Since many patchable electronic music systems are set up using exposed (albeit low signal level voltage) connectors, it is essential that power connectors be kept well covered. There are two reasons for this. First, we must be able to assure the musician that there really is no danger in handling the exposed connectors. Certainly a serious

accident would destroy a lot of credibility on this count even if no lasting damage to the system of the user occurred. The second reason for care is that in promoting the idea that the devices are very safe, carelessness may be invited. Surely most people know standard AC connectors and know they can be dangerous if improperly used. Yet people are careless, and pressed for time in a studio, strange things can happen. I know of at least one case where a plastic power plug was stepped on and crushed. One user took the two prongs and somehow managed to get them individually into the socket. The worse part is that he left them that way for the next user, who fortunately spotted it in time. For this reason, there should be enough sockets added so that no one has to "rig up" anything. And they should all be good three wire systems with proper grounds, or equipment should be otherwise grounded (as in a standard equipment rack). It is my feeling that the user of a system should have the assurance that all metal panels and exposed surfaces are grounded. Take some time and effort to set up the power properly and difficulty will be avoided.

Once you have decided on a layout and have set the various pieces of equipment in their proper places and provided safe and adequate power, the next step is to consider the means of interconnection. All systems will need at least some patch cords even if you don't have a patchable system. There will always be mixers, amplifiers, tape recorders and the like to be connected up. Probably there is no part of an electronic music system other than the patch cords about which nearly everyone has an interesting story. A few examples:

"The Availability of a Particular Patchcord is Inversely Proportional to its Need" [Allen Strange in Electronic Music, Wm. Brown (1972)]

"Necks Are For Wires" [Attributed to Bob Moog]

Some advocates of patch cords over matrix pins or switches have suggested that the matrix systems be supplemented with lengths of yarn to show how the signal actually flows from module to module, and further that copper wires then be placed in the center of the yarn to actually carry the signals!

So there is much that has been said about patch cords. We can add here that even if you don't want to use any, make sure they are good working ones that you don't use. There is no room in the business for bad cords. We suggest that if possible you buy the good (expensive!) ones with the plugs molded on, or at least use the connecting procedure we have suggested in the Musical Engineer's Handbook (Chapter 8b) as this has proven reliable. Another point is that a good place to look for an extra patch cord is in the middle of a patch where one was left from a previous patch, and that a good place to put a patch cord you don't need for a moment is into an unused multiple. There is really no good way to handle patch cords other than around your neck as suggested above, but a permanent solution along these lines causes conversation at dinner parties! We have previously mentioned the idea of using a slotted wooden rack to hold cords along a wall for example. Quite frankly, other common places for cords are in front of the synthesizer on a table or on the floor. This is really not that bad of an idea, as long as they are not in an area where they are kicked around or tangled up. Possible you might want to make a table with two layers where the cords can be stored in the space between layers. One last idea that might be useful is that if you are making up a bunch of new patch cords, make every one a slightly different length. This means that when you find one cord that is just a little short, you know that there is one somewhere that is just right. If you have already used it, check and see if it was not just a little long where you used it. Then you can exchange the two and get a neater patch all around. If on the other hand you have only standard lengths, when you find a short cord, all others will be too short until you jump to the next larger size, which will be at least a little too long.

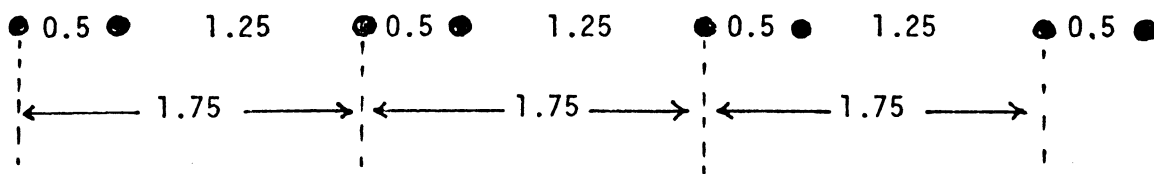
There are of course several different types of connectors that are used in electronic music systems. We, and many others, prefer the standard 1/4" shank plugs as the standard item. These are a little bulky, but are very reliable when properly

used. There will be at least some RCA (Phono) plugs in the system for connecting tape recorders, and possibly also microphone plugs, and a variety of others. The question is then of having enough patch cords so that a variety of terminations are available, or of providing some sort of patching and interconnect panel. We tend to prefer the ideal of an interconnect panel so that all patch cords have the same terminations. This means that there can be fewer cords, and you don't have to worry about pulling out a cord from one end only to find some other, unexpected, and unusable connector on the other end. Naturally, as you pull out more and more useful ones, the probability of locating a useless one goes up. It happens every time (eventually). It is actually quite easy to rig up a patch panel. You just mount a panel full of different types of jacks and start wiring different combinations together. In some cases, you may just want to wire a cable to the back of the jack and run it to the equipment with the strange connector. As an example, consider the problem of running from a synthesizer output to a tape recorder with an RCA input. If you were using patch cords, you would need one with a 1/4" phone plug on one end and a RCA Phono plug on the other. With a patch or interconnect panel, you run a cable from a 1/4" phone jack on the panel to a RCA Phono plug which goes to the tape recorder. You can then think of the result as an extension of the tape recorder input brought up close by and fitted with your standard connector. Generally this is more convenient.

The actual material and size of equipment panels is a matter of personal needs and taste. We generally prefer to build on standard 19 inch wide "Relay Rack" panels as these are standard in many professional setups and work well. The corresponding racks hold the 19 inch panels, are about 6 to 7 feet tall, and cost around 60 dollars new. However, you may be lucky enough to pick up a used one or you can make something that will work out of metal channels or wood. Probably if you have never seen one of these racks, or even if you have and not examined it closely, you will have a hard time seeing just how the holes should be drilled in the mounts so as to accept all possible combinations of different height rack panels. The different heights are:

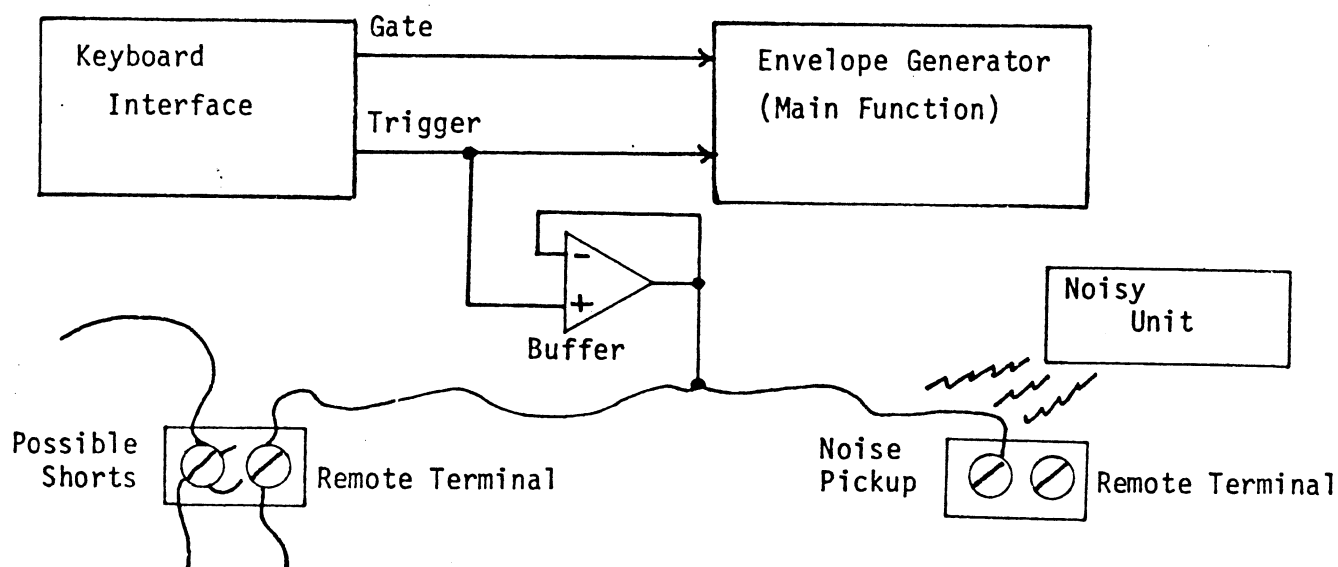
| <u>Height (Inches)</u> | <u>Approx. Cost (Singles)</u> |
|------------------------|-------------------------------|
| 1-3/4                  | \$2.50                        |
| 3-1/2                  | 4.25                          |
| 5-1/4                  | 5.50                          |
| 7                      | 6.75                          |
| 8-3/4                  | 7.85                          |
| 10-1/2                 | 10.20                         |
| 12-1/4                 | 11.70                         |
| 14                     | 13.20                         |
| 15-3/4                 | 14.55                         |
| 17-1/2                 | 15.95                         |
| 19-1/4                 | 17.35                         |

Thus you can see that each panel is a multiple of 1.75 inches high. It might be thought that since the panels are all a multiple of a unit height that holes for mounting them should be spaced equally. This is not the actual case, because the holes in the panels are apparently spaced for the most advantageous mechanical strength. The actual racks have holes that are spaced according to the scheme below:



It is a good idea to mark out this standard spacing even if you are using only one size panel, as you may want to change something in the future.

There are several voltages and signals that tend to be common to at least several modules in a synthesizer system. These of course include the power supply lines (such as +15, +5, and -15), ground, and other main control signals such as pitch level voltage, gates, and triggers. It becomes a question as to whether you should always patch these in, or if they should be hard wired and made switched options. In the case of the power supply voltages, these are obviously hard wired or otherwise attached on a semi-permanent basis with spade lugs or some type of multiconnector socket. The other signals are less obvious. Probably in a keyboard system, we expect to run the control voltage from the synthesizer to all VCO's and all VCF's, but we do not always want these controls in effect, so it is useful to have switches on the driven modules to activate or deactivate this main pitch control. In the case of the gates and triggers, these are mainly used to drive an envelope generator, and are probably hard wired to the envelope generators. Yet they may go to other units as well. For example, it is common to use a trigger to ring a high-Q filter. But probably this is not used enough to make it a switched option. This is easy to patch up if we make sure we include an output jack on the keyboard interface unit. If some signals are used mainly for one function that is hard wired, but you also want to distribute them in various other places in the rack, it might be a good idea to buffer them after the main function but before they go out for additional distribution. This is easily done with an op-amp follower for example. The advantage of the buffering is that a failure such as a short in an external unit will not defeat the main function. In addition, noise picked up on the external lines will not get back in to possibly confuse the main function. An example showing the distribution of a trigger signal is shown below:



There are many other problems you may run into when trying to set up a system. It is best not to be in too much of a hurry, and to do things carefully. Other things that are useful are to keep a good notebook on the system, and to have a regular maintenance schedule. While there is usually little that can go wrong with a system once it is set up, a regular time set aside for maintenance will give you time to do some new things, and to allow you to check over the system and perhaps again look at some things you have forgotten and may want to use.

## 2-3 HOW I MAKE MY P.C. BOARDS IN A MATTER OF A FEW HOURS:

-by Bernie Hutchins

Over the years I have developed a simple and reliable method of producing P.C. boards which I find so logical that I really don't see how anyone else would want to try anything else for small quantities. Although I have described this in the MEH and in the Application Notes (AN-14 to AN-18), it seems to be useful to go over it briefly again here in the newsletter. We get a lot of questions about boards, and those builders who I have converted to my system will vouch for it. Typically you should be able to prepare a board (a VCO say), etch it, and begin soldering on the parts within a couple of hours of the time you first decide on which schematic to use.

Some essential points about the method: (1) You are going to be working on only one side of the board - the copper traces and all components will be on the same side. You will only drill holes in the corner to mount the board. (2) You will apply the resist pattern using a fine artist paint brush and a thinned down (with lacquer thinner - add about 50% to the original volume) lacquer paint. (3) You will be making up your layout as you paint, and painting freehand with a minimum of guidelines. You will be working directly from the schematic to the final board with no sketch in between. While all this may seem like it requires immense concentration and an artists hand, you should not have too many problems if you just accept the fact that it is not going to look 100% professional, and you are going to have a few more jumper wires than you might need if you worked on the design for a few days!

The first step is to select a board of the proper size. This is a matter of placing the IC's and the larger components in approximate positions using the schematic layout as a basic guide. Next clean the board with steel wool pads or scouring powder using a circular motion to assure that scratches are randomly placed making it easy to see pencil lines later. Next I make a border around the edge either with a ruler (to be painted later) or with black electrical tape if I can stand a wider edge. This serves as a ground area. Then lay out the IC's and other large components again. Try to line IC's lengthwise so that power supply lines can be run underneath. Make a pencil trace using a ruler for all supply lines to all IC's. Then accurately position the IC's relative to the supply lines and mark the positions where their "feet" contact the copper with a little dashed pencil line.

The next step is the most important one - you have to start painting. Plunge right in! If you think about it, you have already arranged to mount all the IC's at this point and have arranged for them to get power. All that remains is to connect up all the smaller resistors, capacitors, and other components, which after all is the really interesting part of the circuit anyway. Happily, most of these components are "free jumpers" - that is, they naturally will jump over supply lines and other lines which you add. Somewhere you will probably run into trouble and leave out some connection. Just leave a little "tab" of copper at each end, and remember to run a wire later when you solder on the components. After you have made a board or two, you will find this painting to be much easier. You will find that certain structures (around op-amps for example) are so common that when you see an op-amp on the schematic you will not even have to think about how you are going to paint them. It will always be the same. One especially useful result of this sort of board preparation is that everything is on the same side of the board and is easy to see. Op-amps are mounted as they are usually drawn. You can almost read the board just as you would a schematic, making it easy to check and troubleshoot as well.

Once the board is painted and dry (with the thinned down paint, this is just a matter of a few minutes), you will be ready to etch it. I use ordinary ferric chloride etchant (available at Radio Shack stores or at similar dealers) and dump the

whole bottle into one of those plastic freezer dishes (rectangular, about 6" by 8" with a cover) which is used for etching, and storage of the solution for later boards. I find the best way to etch is to just float the board and come back later. I don't heat the etchant or agitate the solution during etching. It may come as a surprise to some readers that you can float a board on top of the etching solution but all this wonder can be attributed to surface tension of the liquid, and the relatively dense solution (which I find decreases as the etchant is used). By floating the board, I find there is no trouble achieving a good etch within 15 minutes to 1 hour depending on how far you want to push the useful life of the solution. There are two things to watch out for. Be sure when you float the board that you do not get an air bubble under it or you have a spot that does not get etched the first time. One way to prevent this is to tilt the board ever so slightly as it makes contact with the etchant, and then lower it slowly. The other thing is that you should make sure the board does not sink or else it just sits on the bottom with a thin layer of spent etchant against the copper and you have to etch again. Once the board goes down, it will continue to do so unless you remove it and dry it thoroughly before trying to float it again. Why not let it set face up in the bottom of the dish? Well, you can do this with a fresh solution, but with a used solution a sort of "mud" will settle on the copper surface and tend to inhibit the etching process. Also, it is harder to remove a board from the bottom of the dish than from the top. You will probably get some etchant on your hands as you work and this will not hurt you, although you should wash it off. The biggest problem from the ferric chloride solution is that it tends to stain any porous surface (wood, fingers, cloth) so take proper precautions.

It will be useful to go through a brief example. We will show some sketches of the entire process, but for a very small circuit. Fig. 1 shows the schematic of a simple two op-amp triangle-square oscillator. Normally, this circuit would be part of a larger circuit and would not get its own circuit board, but we want to keep things simple here so that the example remains clear. In Fig. 2, we have selected a circuit board, painted a border around it (for mounting and ground) and have lined in and painted the power supply lines (the positions the op-amps will take are shown with dotted lines). The pins other than the power supply lines are just marked with pencil dots at this point. Fig. 3 shows the remaining paint lines which are put in freehand. The board is ready to etch at this point. Fig. 4 shows the board after etching, cleaning, and mounting of components. Note that two jumper wires are used, and that the components are shown with long leads while in practice you would keep them as short as possible.

The example should make the general ideas clear. In addition to the idea that this is really a simple thing to do, I wish especially to point out the striking similarity between Fig. 1 and Fig. 4, the original schematic and the final circuit board. In fact, you might think of Fig. 4 as a "conducting schematic." I believe you will find that many of our schematics will easily take on a corresponding circuit board form. Readers who may have been sold on this general idea may want to take a look at more of the suggestions in the MEH and in the application notes listed above (these application notes are part of the "preferred circuit collection" as it is currently being offered).

Other readers may be thinking, "Yes, it's fast and logical, but not neat." If it is necessary to be especially neat, it is possible to improve the technique so that things look better. However, we insist that only one side of the board be used, so if your idea of a neat professional look is a board that looks like one in a commercially made unit, this will not do the trick. Assuming you just want your own equipment to look a little better when you open it up, but in general want to follow this method, we suggest two improvements. First, after painting on all the etch lines, but before etching, take a few moments to neaten things up. This can be done with the paint brush when it is desirable to increase the width of some lines or to neaten up a corner, or with a sharp pointy object for removing minute amounts of resist paint. A good tool here is made from a 1/4" to 3/8" wooden dowel with a small wire brad nail driven in the end. You then file or grind off the head to a sharp point, effectively making

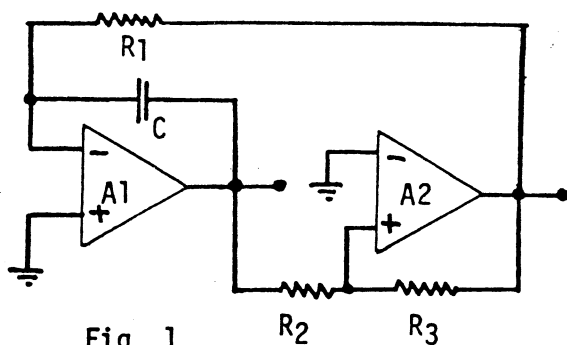


Fig. 1

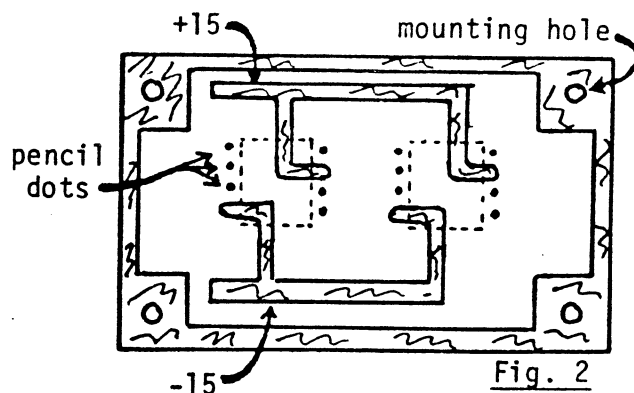


Fig. 2

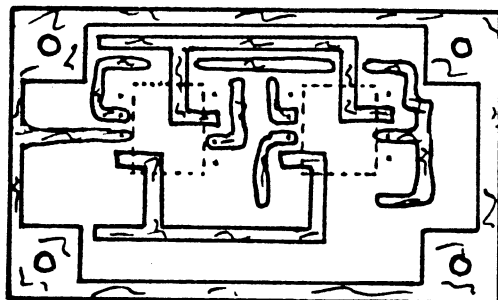


Fig. 3

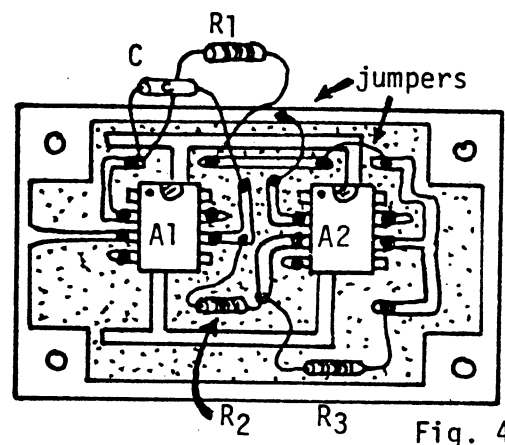


Fig. 4



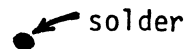
Resist  
Paint



Copper



Insulating  
Board  
(etched)



solder

yourself a "pencil" with iron lead. You can use this freehand to scratch off bits of resist paint (and to "erase" mistakes during the original painting), or you can lay a ruler along a painted line and run the point up and down the edges removing any slight imperfections. Five minutes work here will greatly improve the appearance in many cases. The second improvement is to tin the board. This is often a good idea in any case since it lowers the resistance of the lines and covers any hair-line cracks, as well as making the actual mounting of components easier. To do this, you first clean the etched board thoroughly, and possibly even buff it slightly with toothpaste. This makes it easy to apply a light coat of solder to all copper surfaces. The board is now tinned, in a satisfactory but obvious way. For a more professional look, you might want to try for a "plated" appearance. To do this, I turn the board upside down and remove excess solder by letting it run down on to the hot soldering iron. This removes much of the excess solder and leaves a shiny appearance with possibly grainy looking areas and bits of burned rosin flux. The final trick now is to use steel wool on this board to even the surface, remove the excess shine, and to remove the burned flux. The result is an evenly tinned appearance and a board appearing clean of any excess solder or flux. The components are now mounted in the normal manner, just as they would have been on the bare copper if we had chosen not to tin the board.

I hope this all convinces you that you can, if you wish, do your own boards yourself, and that obtaining boards need not be a major problem. Keep in mind that as long as you are at least partly in the "build it yourself" business, making your own boards is consistent with the rest of what you are doing. In addition to the advantages that have been described above, we should also point out that the preparation of a board is one additional step that helps you to become familiar with the circuit you are building. Since, as we pointed out, the method above is very similar to a redraw of the schematic, preparing your own boards in this way will force you to become aware of all the components in the system and their interconnection.

## PUTTING YOUR CIRCUIT BOARDS INSIDE SOMETHING:\* -by Bernie Hutchins

Many of us get used to building circuit boards, bench testing them with some sort of combination of alligator clips, jumpers, etc., and then going on to another project. After a few modules, you run out of bench space and have to consider putting things into some other sort of enclosure. This is not difficult to do - just difficult to make up your mind about what to do and when to do it.

You probably have in your mind some sort of physical setup in which you see your modules all ready to go - probably in the form of some portable synthesizer you have seen or used, or in the form of some large studio setup. You may also have discovered that the packaging that you desire will cost you the same amount of time and money that the actual electronics has so far. Naturally, this can be very disappointing news, and you may decide to set your sights lower. Fortunately, many things that do not look quite as good will serve just as well. Let's look at an example.

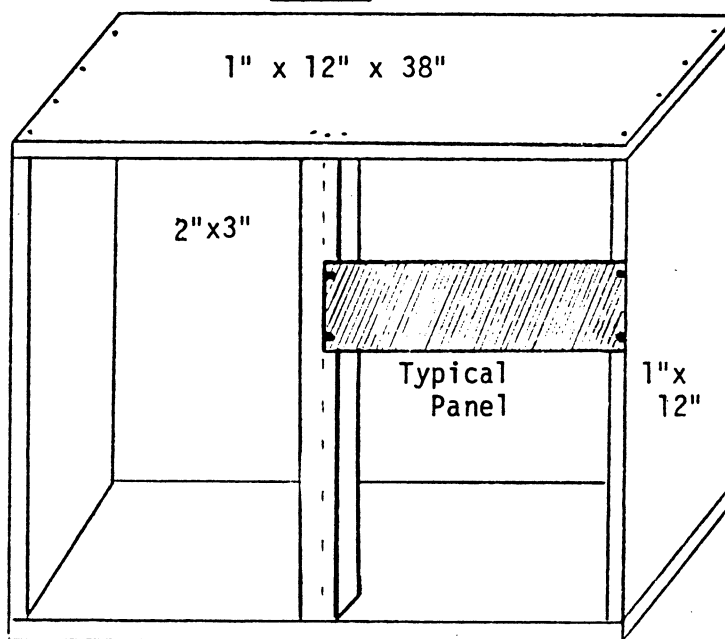
As we often do, we will assume that you decide to build on standard 19" "rack panels" which come in a variety of widths [see "Hints on Setting Up an Electronic Music Working Area," EN#82 (17)]. These fit nicely on a standard "relay rack" which is great if you have one, but which otherwise costs at least the better part of \$100. Many readers might like to consider the much less expensive wooden rack of the general form shown in Fig. 1. I use one

like this in addition to a standard rack (which I got as surplus at an action for about \$20 - watch for this sort of thing), and a bench setup. Note that the physical setup is similar to a medium size studio unit. The whole thing is in fact basically a wooden box made of 1" lumber that is 10" or 12" wide. The box is 38" wide on the outside (the length of two side-by-side 19" panels). A center support wider than 1" lumber is added (2" by 3" stud suggested). It is desirable to make the box rigid with angle irons, or by installing a sheet of thin plywood on the back (assuming you don't want to get in the back

anyway. Alternatively, relatively modest bracing will suffice because once a fair number of panels have been securely screwed into place with heavy flat head wood screws, the whole structure becomes very rigid - like it or not. Well, that's the general idea. You may want to do a lot more with the idea. At least we now have a place to mount our panels, and if we have made the circuit boards, we now have one logical step remaining.

In order to attach the circuit boards to the panels it is necessary to mount all jacks and controls on the panel, and then use some suitable means of securing the board to the panel, and finally the interconnecting wires are soldered. The exact

Fig. 1

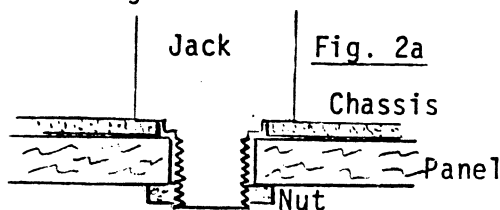


\*See "How I Make My P.C. Boards in a Matter of Hours," EN#90 (3) for information on getting to this point.

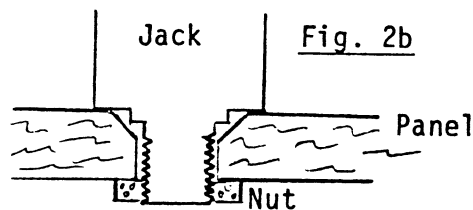


layout of the panel is a matter of personal taste. Some planning of this should be done however, and it is probably a good idea to arrange the jacks and controls on top of the undrilled panel to make sure there is enough room. To drill the panel, you will need an electric drill and several bits, and a round "rat tail" file. Most pots and standard 1/4" phone jacks require a 3/8" mounting hole. If you are using an aluminum panel, you can drill first 1/4" holes with a 1/4" bit in a drill that is intended for a maximum drill shank of 1/4". You can drill these holes larger with a 3/8" drill that has a 1/4" shank, but don't try this oversized drill unless you have already drilled out to 1/4" and are using aluminum panel material. In an emergency, you will be surprised how easy it is to enlarge a 1/4" hole to 3/8" with a good round file. This may save you a trip to the hardware store, but you will eventually need the 3/8" bit. If you have better drilling equipment than we have suggested above, so much the better.

After drilling, there will be a bit of a burr around the edges of the holes. Leaving this burr will often result in a better electrical connection for jacks, so you might want to leave it, but it is a good idea to file it down, or to scrape it off with a pocket knife for other mountings. If you are going to mount jacks or controls with a short bushing, or if you are going to have to mount controls through another layer of metal (such as a chassis against the panel), you will have to get the burr off completely. If you don't, there may be space left between the layers and you won't be able to get bushings through far enough to get the nut on. For some components (such as jacks), you may have to leave off a washer, or even file or drill a space so that the item to be mounted can be set in further (see Fig. 2 below). Keep the round file handy also for correcting a poor alignment between a panel and a chassis when a jack or control binds as you try to push it through the mounting holes.



Setting Jack In by Using  
1/2" Hole Through Chassis



Setting Jack In by Filing  
Bevel Around Hole in Thick Panel

Once all jacks and controls are mounted, you will have to secure the circuit board in place. Because electronic music devices have so many inputs, outputs, and controls, it is necessary to reserve as much space as possible for controls, and to use as little as possible for mounting the boards. Another consideration is that if you intend to use the unit out of your "rack" or if you must protect it from dust, you may have to enclose the unit eventually. Two extreme cases are shown in Fig. 3a and Fig. 3b. In Fig. 3a, a chassis is mounted on the back of the panel and all controls and jacks pass through both (see Fig. 2a). Note that here the chassis is upside-down from the more usual orientation, but this is preferred here since it provides a sort of "box" to mount things in, and you can work on the unit at your bench while the panel is right side up and facing you. When everything is finished and working, you can mount a cover (a chassis "bottom" or just a sheet of sheet aluminum) and the whole unit is enclosed. Fig. 3b shows a mounting that is somewhat fragile, but which is dirt-cheap and can often be used where the unit is to be mounted on a permanent basis in an enclosed rack. Here we have simply soldered the grounded copper edge of the circuit board to the ground tabs of some convenient jacks. This in itself is a little risky, but if you use solid hookup wire running from points on the board to jacks and controls, and keep these wires straight, you will be surprised how solid the whole unit will become. Some other techniques can be suggested here, and probably the reader already can think of additional ideas along these lines. Fig. 3c shows how a length of aluminum angle bracket can be used to securely mount a board to the panel. This works well where there is one main board in the system, and where

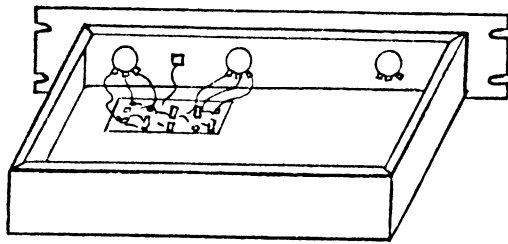


Fig. 3a Use of Chassis Box

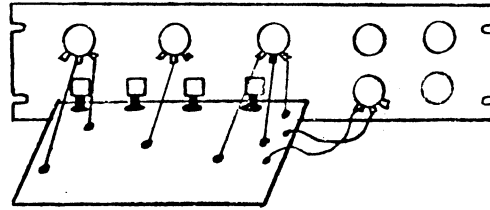


Fig. 3b Direct Solder Mounting

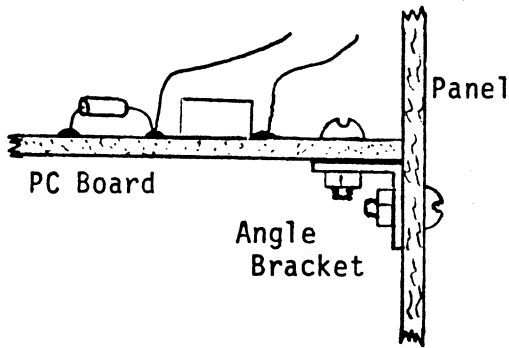


Fig. 3c Angle Bracket Mounting

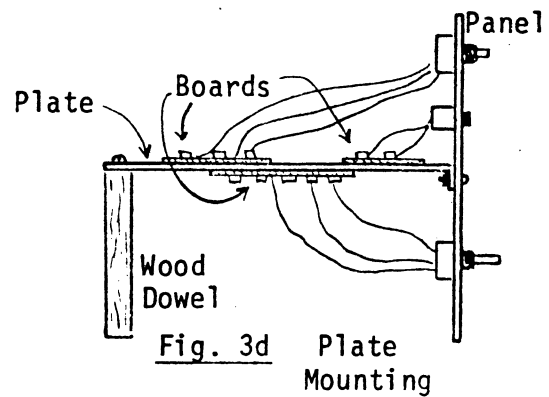


Fig. 3d Plate Mounting

there is some extra panel space available for the bracket. Fig. 3d shows a method that is well suited to a requirement where a large panel is needed for a variety of relatively small modules on separate boards. Basically one leaves a center ridge along the length of the panel and mounts the jacks and controls to either side of this ridge. An aluminum plate is mounted on the back side of this ridge perpendicular to the panel as shown, and the circuit boards are fastened to the plate on both sides. One or more lengths of wooden dowel attached to the back of the plate with wood screws are useful for supporting the assembly on the test bench.

Here we have given a few ideas and hope we have left the impression that things can be relatively easy if you want them to be. The exact details of your own setup will depend on your own needs and resources.

\* \* \* \* \*

Fabricating attractive front panels for a synthesizer can be a real challenge at times, and so I want to describe a system I have evolved over the last couple of years. You will note that I call this a "system", for it is only by some systematic approach that consistent results can be expected. I have built some thirty modules using this scheme, and so have had quite a bit of time to learn from my mistakes. I hope then, that this article will save you from some of the pitfalls I have encountered.

I set fairly high standards for attractiveness when it comes to front panel design. Now, I know there are probably some readers out there who don't feel this strongly about the esthetics of front panels, but nonetheless this article may have a few tips for you too.

What I am going to reveal then, is a method for taking ordinary blank rack panels and turning them into front panels for synthesizer modules. The end result will be quite attractive, and could easily be mistaken for a "professional" job.

Perhaps the first question to consider is where to obtain materials. Well, we will need a blank rack panel, and they are fairly common. Places such as Burstein-Applebee and Allied Electronics are good places to find these. I think the prices might alarm you though! For example, a blank, double width panel goes for around five to eight dollars! (All rack panels are 19" wide; the height comes in multiples of 1 3/4". A single width is 1 3/4", a double width is 3 1/2", a triple width is 5 1/4", and so on.) Well, I don't know about you, but to me that seems kind of a rip-off; I mean, who ever heard of eight dollars for a little, skinny piece of aluminum?

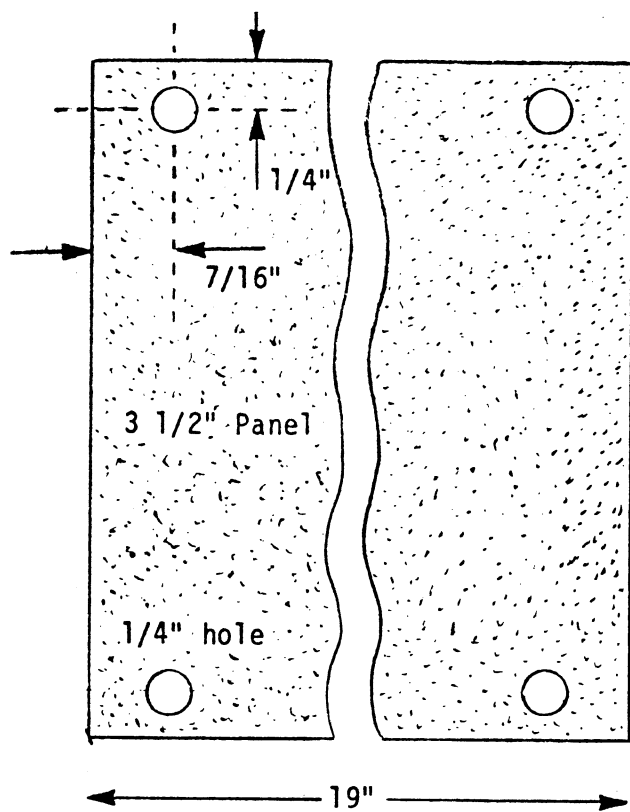
I decided to check the alternatives, since I didn't want to play my home-brew synthesizer in the poor-house. I discovered the following interesting fact. Local sheet metal houses will cut you the right sized panels for anywhere from one-fifth to one-tenth the prices mentioned above. The place where I go will cut me single width panels for 65¢, double width for \$1.30, and so on. Quite a savings!

Of course you will have to cut your own mounting holes, but for that kind of price break it's worth it. (It takes me about ten minutes to cut the four notched mounting holes.) When you go to the sheet metal firm, be sure to tell them that you want 1/8" aluminum stock, 19" wide by whatever panel height you have in mind. And you will probably make the man happier if you order more than one panel at a time, since he can then cut a big strip 19" wide, and then sever it down from there to the individual panels.

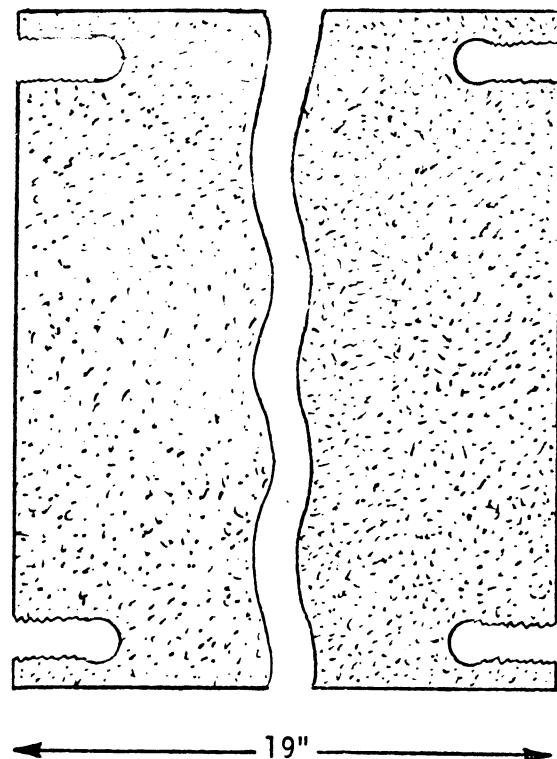
You may get a choice of quality of stock, in which case go for the least expensive (i.e., the worse looking), since in a later step we will be removing all the surface blemishes anyway.

Now a word about making mounting slots. Figures one and two give the basic ideas. In Figure one, four holes have been drilled or punched in the corners. Their diameter should be 1/4". An easy way to locate their positions is to use a standard store bought panel as a template. Or you can take the measurements off the rack case, or use some of the standard positions indicated in Figure three. (Figure four shows the standard spacing of holes for a rack case. The spacing is such that any two panels of any size may be positioned together.) Now, with a hack saw slots are cut into the holes as shown in Figure two. After this, a flat file and a rat-tailed file are used to smooth up the edges of the cuts.

After following the above steps, you will have a standard rack panel for a fraction of the normal cost. "The only trouble now", you say, "is that it looks ugly, all those nicks and scrapes!" Well, I told you not to worry. Simply load an electric sander with some fine sandpaper and go to it. Don't be concerned about the sandpaper "design" being etched into the panel, since they will be covered later by



**Fig. 1** Panel drilled with 1/4" holes. Spacing of holes is always 1/4" from edge or 1.5" from edge (See Fig. 3)



**Fig. 2** Slots are rough sawed out. Cut slightly inside and smoothen cut edges with flat file.

some paint primer. But go at those nicks and scrapes, and after several minutes sanding you will have a smooth panel

The next step is to buff the panel with some 00 gauge steel wool. This will do wonders to the surface and will give it a smooth and silky feel. After buffing, wash the whole panel with soap and water to remove any grit and steel wool particles. Wipe dry with a lint free cloth.

Now we are going to put a temporary "cover" on the panel as an aid to the layout process. Using strips of masking tape, cover the entire panel front surface (decide which side you want to use for the front first). This step is easier if you use two inch wide masking tape. The general idea is to cover the front surface with a protective layer which will accept pencil markings.

Working from a life-sized mock up, prepared earlier (more about this later), mark the various spots on the panel (on the tape) where you will need to drill holes. Then with a metal punch and hammer, indent these various centers which you have just marked.

At this point you can drill or punch the holes in the front panel. If you use a drill be sure to make some pilot holes first, and in general be very careful since that metal can really cut if the drill bit should snag and start the whole panel spinning around. I always clamp the panel to my workbench on top of a piece of scrap lumber and this holds the panel firmly in place.

After the holes have been made you can remove the tape. We are almost ready to paint the panel, but first you should clean the surface with a cotton ball and some alcohol. The alcohol serves to clean off any soap film from an earlier step, and will also clean off any adhesive left from the masking tape.

We can now paint the panel with some primer grey. Don't neglect the primer, since it has the property of smoothing out or filling in any small abrasions.

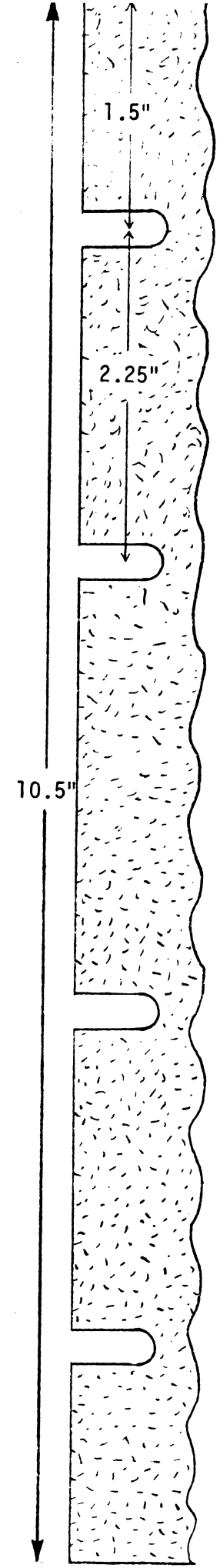
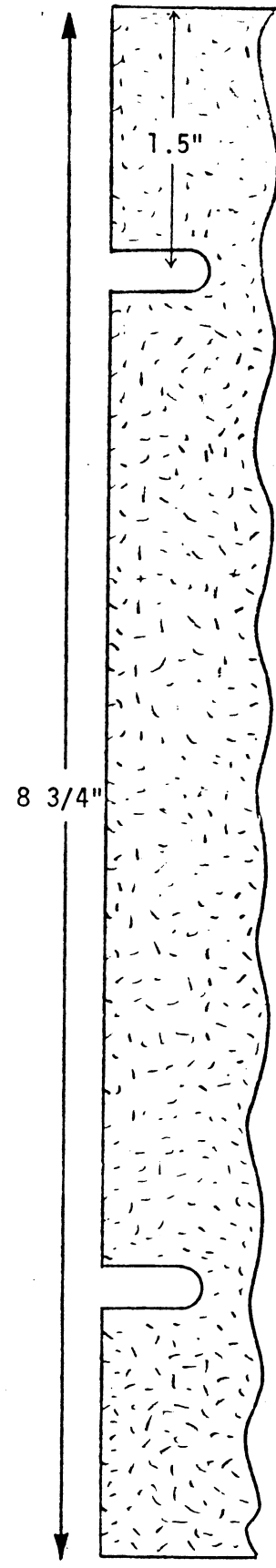
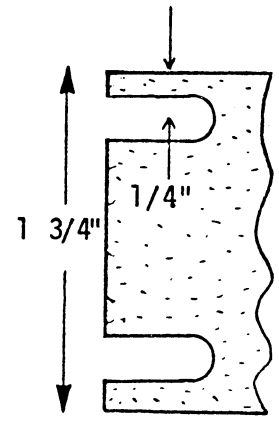
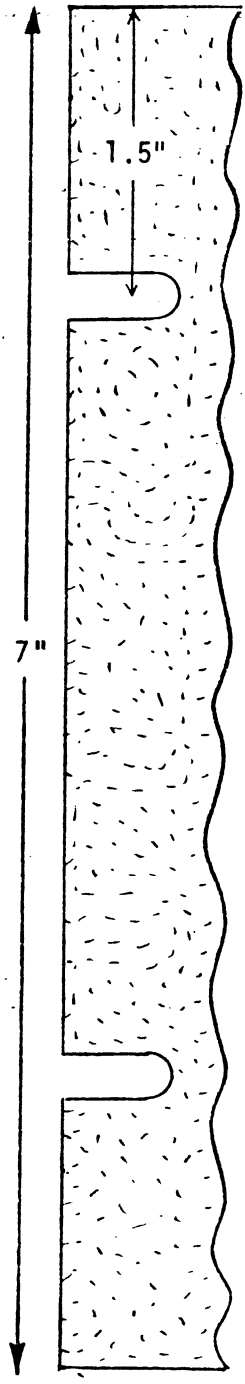
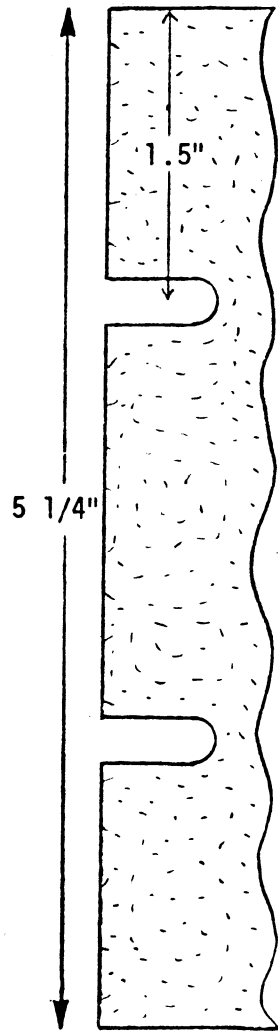
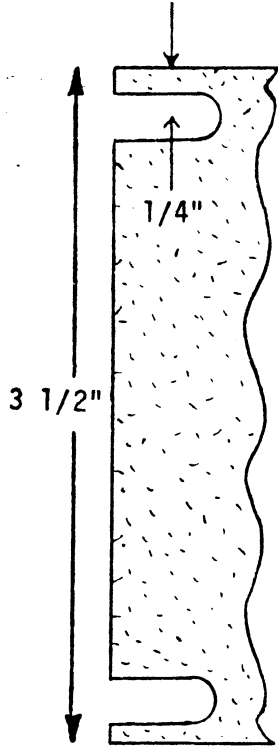


Fig. 3

Standard Rack Panel  
Sizes and Mounting Holes

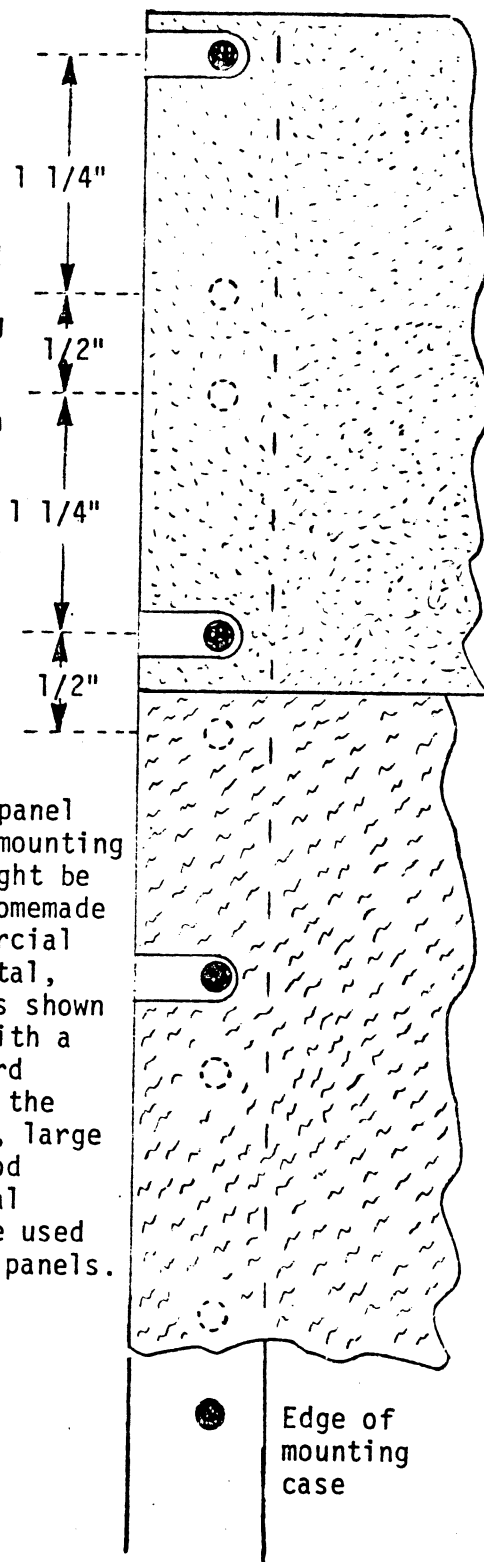
The end result will be very smooth and silky to the feel. When the primer is dry, paint the panel with any color spray paint that you like. Epoxy spray paints give an extra hard surface and are well worth the extra cost. I use a pale yellow color since the dry transfer letters I use show up very nicely on it. Black is also a good color if you happen to have access to white transfer letters.

At this point, you can label the panel. Dry transfer letters give a very nice touch and are really quite easy to use. If you make a mistake, simply lay a piece of masking tape on the error and lift it up, and the letters will lift up on it. In the case of VCO's, you may want to do what I do and actually draw the output waveforms below their respective jacks. I use a special pen for this. The pen is an alcohol base affair, with a narrow tip, and it is available at Radio Shack. The intended use for the pen is resist ink work (for printed circuit boards). In addition to the pen I use a standard schematic template to generate the various shapes. Let me tell you, the use of the template and that special pen has really classed up my modules, and it's really impossible to tell that the various symbols were created by hand. There are all sorts of tricks you can pull to create graphics and you will discover more as you go along. When you are done laying out the panel to your satisfaction, spray paint the surface with clear plastic or lacquer. Allow this to dry and spray again. And one more time, spray again. The purpose, of course, is to create a very hard protective surface on top of those relatively fragile dry transfers.

Dispite this protection, you should take extra care when mounting jacks, pots, and the other controls. A slip of a pliers or wrench while tightening a nut can easily mess up your hard work.

At this point you are done with the panel and the results will be durable and attractive. I think you will be surprised at how the panel no longer betrays its heritage; no one will ever know that it was home made and that it started out as a piece of ordinary aluminum.

**Fig. 4**  
A portion of the left side of the panel mounting case showing a 3 1/2" panel mounted with a larger panel below. Note how the sequence of holes spaced at alternating 1 1/4" and 1/2" spacings allows for any sequence of panel sizes. The mounting case edge might be wood for a homemade case. Commercial racks are metal, and the holes shown are tapped with a 10-32 standard thread. If the case is wood, large flat head wood or sheet metal screws can be used to mount the panels.



I mentioned above that you should lay out the panel from a mock-up. It is essential to prepare for this by doing a full-scale drawing of how you want the panel to appear. By doing it full scale, you can play around with the actual pots, jacks, and switches themselves to see how they fit together. It has been my experience that this mock-up is the most important part of the whole thing.

There you have it, then, my guaranteed method for making panels. It may seem like a lot of work, but you will find that if you have a clear idea of what's involved, you can actually go through many of the steps without thinking! The night before I ever build anything, as I lay down to sleep, I visualize in my mind just what I will be doing the next morning. I find that this mental preparation makes the actual work go very fast, since I am essentially going through the motions. After all, I've already built the thing (in my mind) so what's so hard about building it again? Who'd have ever thought "Psycho-Cybernetics" would ever come to electronic music?

The point then, is to develop a method. I hope this article will help you to develop your own method, and I hope your front panels will please you as much as mine have pleased me. Happy building!

\* \* \* \* \*





## 2-6 SOME MORE SUGGESTIONS ON ACTUALLY GETTING CONSTRUCTION UNDERWAY:

-by Bernie Hutchins

Let's suppose that you have purchased all the parts you need for an electronic project, have read how to make circuit boards (EN#90) and how to package your finished boards (EN#98). You are still not ready to build, or at least you need to check out certain supplies and tools, and to set up a reasonable work area. Here we will offer some suggestions along these lines.

### CHECKLIST OF TOOLS AND SUPPLIES

Unfortunately, it is not at all unusual to have a project nearly complete only to realize that you need a tool not readily at hand (such as a hex key wrench for a set screw that you had been thinking was a standard screwdriver slot). Often the tool is "somewhere around here!" but not in your hand. At still other times, it would be nice if you had a \_\_\_\_\_ so that you could \_\_\_\_\_ but you just do without. Perhaps to make things easier, and to achieve a better psychological state through thorough preparation, it is useful to make a list of the things you will need, and then make sure you have them all gathered together, or pick them up at the next opportunity. Below we make a "checklist" of things you need, tell you why you need them, tell you about what they will cost, and then suggest where you can get them.

| <u>ITEM</u>          | <u>WHY YOU NEED IT</u>        | <u>HOW MUCH?</u>         | <u>WHERE?</u>                                                               |
|----------------------|-------------------------------|--------------------------|-----------------------------------------------------------------------------|
| Soldering Iron       | ---                           | About \$3                | Hardware section of Dept. store or electronics supply                       |
| Solder               | ---                           | \$6-\$10 per 1 Lb. Spool | Good hardware store or electronics supply place.                            |
| PC Board             | ---                           | ~\$1/board               | Mail order electronics, Radio Shack (etc.), and watch out for free "scraps" |
| Etchant              | ---                           | \$2/bottle               | Radio Shack, etc.                                                           |
| Etchant Tray         | ---                           | about \$1                | Plastic "freezer" container from Dept. store                                |
| Electric Drill       | all ordinary holes            | about \$10               | Discount Store                                                              |
| Drill Bits           | ---                           | varies                   | Hardware or Dept. stores                                                    |
| Oversized 3/8" bit   | pots and jacks                | about \$3                | Hardware store                                                              |
| Round File           | enlarging and deburring holes | about \$1.50             | Dept. Store Hardware                                                        |
| Flat File            | Soldering iron care           | about \$1.00             | Dept. Store Hardware                                                        |
| Tin Snips            | Cutting ckt. board            | about \$5                | Dept. Store Hardware                                                        |
| Medium Screwdriver   | most things                   | about 70¢                | Dept. Store Hardware                                                        |
| Small Screwdriver    | knobs                         | about 40¢                | Dept. Store Hardware                                                        |
| Phillips Screwdriver | maintenance                   | about 70¢                | Dept. Store Hardware                                                        |

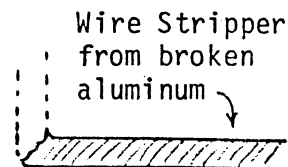
|                                |                                                                                  |           |                                                    |
|--------------------------------|----------------------------------------------------------------------------------|-----------|----------------------------------------------------|
| Hardware, Nuts,<br>bolts, etc. | Packaging                                                                        | varies    | Dept. Stores, Hardware<br>Stores, Digi-Key, Scraps |
| Wire                           | Hookup                                                                           | varies    | Electronic supply, Radio<br>Shack etc., surplus    |
| Steel Wool                     | Cleaning PC Boards                                                               | \$1.00    | Grocery store, Brillo, SOS                         |
| Black Tape                     | Large "etch resist"<br>areas on PC Boards                                        | \$0.70    | Hardware or Dept. Store                            |
| Enamel Paint                   | PC Board Resist                                                                  | \$1.50    | Dept. Store                                        |
| Paint Brush                    | Painting PC Boards                                                               | \$0.50    | Dept. Store, Art Store                             |
| Pencil-Type<br>Ink Eraser      | Removing dried paint<br>PC Board "Mistakes"                                      | \$0.30    | Dept Store, Stationary<br>Store                    |
| Sharp-Pointed<br>Knife         | Removing Post-Etching<br>PC Board "Mistakes"                                     | \$2.00    | Dept. Store or<br>Hobby Store                      |
| Epoxy Glue                     | Mounting of some<br>components, insulating<br>of high-voltage<br>terminals, etc. | \$1.00    | Dept. Store                                        |
| Spray Paint                    | Panel Work, etc.                                                                 | \$1.50    | Dept. Store                                        |
| Press-On Letters               | Panel Work                                                                       | \$1/sheet | Stationary Stores                                  |
| Clear Lacquer Spray            | Protection of Press-<br>On Letters                                               | \$1.50    | Dept. Store                                        |
| Electrical Sockets             | Bench Work                                                                       | varies    | Building Supply Store                              |

Probably you can make many additions to this list. Note that the total list is not all that expensive, so it may be worth getting all the things you need or think you will need to save time later.

## SOME USEFUL THINGS THAT WON'T COST YOU MUCH AT ALL

Happily, there are a number of very useful things for construction that won't cost much, if anything. First, we mention that you could add to the list above some sort of pliers. We have not added it because there are many different kinds. You really do not need one that has a wire cutters to it. The most useful wire cutters is an ordinary fingernail clipper; an old one will do. Another type of clip is the alligator clip, usually used for test leads. I never have had a wire stripper, and don't really need a fancy one. An alligator clip works quite well, and one that is not quite right (the teeth do not close exactly) is probably an excellent wire strippers. Just let it bite the wire and insulation, and then pull out, leaving the torn off insulation in the jaws. You can also use the same clip for many of the jobs a pliers would be used for, such as holding parts that get hot while soldering. The alligator clip has the advantage that it grabs on to the part all by itself.

Another useful wire stripper is a piece of broken off aluminum. The diagram at the right will give you the idea. Actually, mine started out as a home made angle bracket, but the aluminum was too thick and it broke, leaving a "burr" or edge that was just right for stripping wire between the edge and your thumb.



Keeping a work area orderly is always a problem (for me at least, it has been impossible) so anything that can help should be considered. One thing that will work well as a "catch-all" for small parts is a small size box of corrugated cardboard cut off all the way around about an inch to 1.5 inches above the bottom.

This box has the outstanding advantage that you can store parts on a temporary basis by sticking them into the corrugation holes exposed by the cutting. If you are really into organization, you can order the parts along the edge in the order in which they are to be installed.

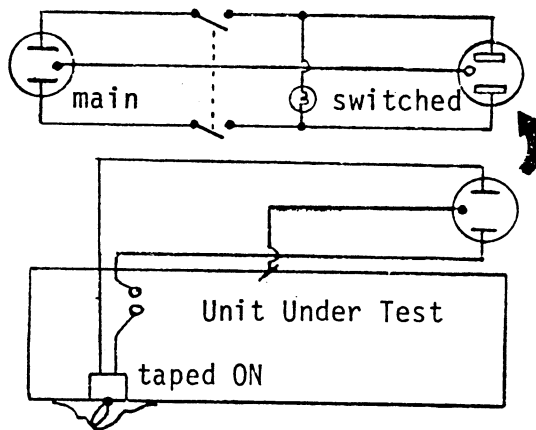
Parts storage is simple if you go out and get a bunch of those parts cabinets with the metal frame and the numerous plastic drawers. The problem is that these are usually in the \$12 to \$18 range. Even one will help a lot, and is definitely worth it, but if money is tight, consider looking for other small boxes of uniform size and makes some sort of cabinet from them. By the way, the cabinets that run from \$12 to \$18 are popular "sale" items in department stores (particularly Montgomery Ward) and can be had at times for as little as \$9.

Don't overlook "junk" electronics as a source of useful parts, not electrical ones necessarily, but hardware and metal parts. Just about any piece of electronic gear can be salvaged to get nuts, bolts, etc. Often, a metal chassis or even a panel with only a few holes can be salvaged. However, be sure to throw out the obviously useless stuff.

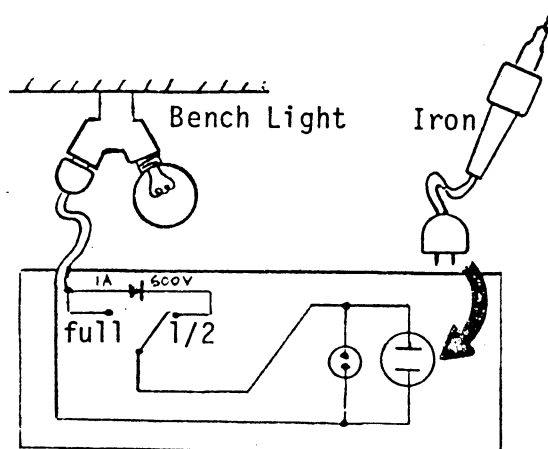
## SETTING UP THE WORK AREA OR WORK BENCH

Probably your projects will take at least several days, so the kitchen table or similar is not the best place to set up. Choose a place you can keep for a while. Keep things in one place if possible, with the exception of the circuit board etching which should be done somewhere where a mess won't matter much. The ordinary type of etchant will stain anything it can soak into. Cloth and wood will be stained and it won't wash out.

In setting up, make sure you have electrical outlets available. However, I suggest two wirings that are non-standard be added for safety reasons. First, there is the problem that the unit you are working on must have some access to AC lines, and secondly, you are using a soldering iron that is hot and a potential fire hazard. These wirings were discussed briefly in response to a reader's question in EN#95, but here it will be useful to give some diagrams. The first system is simply a switched socket, but the switch is a double-pole single-throw type, not the usual SPST type. This can be any panel type toggle switch, or the standard household style switch can be found with DPST features. The important thing is that it cuts off both sides of the line, so it does not matter which side is really the hot one. To use this, turn the switch to the unit you are working on to the ON position, and tape it on. Plug it into the special socket, and now, to cut power, you will have to use the DPST switch (see figure at right).



The second wiring system (shown in the figure at the right) is one for your soldering iron and has two important features. First, it derives its power from the workshop light. The light must be turned on for the iron to be on, and conversely, if the light was turned off, then the iron must have been turned off too, and you need not worry about it. The second feature is a half-power switch, an old trick of inserting a diode in series with the iron, thus blocking half the AC cycle, and reducing the heating capacity of the iron. It has generally



been used in the past as an "idle" feature. Here I suggest you use it as a "normal" mode, with full-power as the option. I generally use the Ungar type "soldering pencil" since it is cheap and works well. The heating elements come in a variety of wattages from about 22 to about 47 watts. The lowest wattage is all you need for most IC work, but for the slightly larger jobs, a larger wattage is essential. You can use about a 42 watt heater for a general purpose iron, but this runs a little hot during idle, and the copper tips burn up fairly fast. By the way, I much prefer the copper tips to the iron clad ones even though the iron ones seem to last much longer. I prefer the copper even though, and probably because, they require periodic filing and reshaping. The system which I have found to be ideal is the 42 watt heater run half-power most of the time, except for initial warm-up and for heavy heat jobs, where the switch is put on full-power. Both the tips and the heating element will last much longer this way, and the more massive metal structure of the 42 watt heater seems to have a better heat reserve and (run half power) seems to work better than the corresponding 22 watt heater. Note that a neon indicator is used. In the half-power (DC) mode, only one electrode of the neon light glows, so it is easy to tell which position the switch is in.

Both these wiring setups should use proper AC wiring techniques. The switches and sockets should be installed inside the proper metal outlet boxes. A visit to a building supply store and a little imagination will allow you to plan your setup so that it is safe and convenient.

## TEST EQUIPMENT

We get a lot of questions about test equipment. Happily, people generally assume that they will need a lot more, and better quality than they actually end up with or even need in the end. Let me tell you what I use.

First, I have a very old scope. It has no particular features except it does display a waveform without any fancy coaxing. While it is not really accurate, it does tell me the things I need to know about the waveshape. I never try to make accurate voltage or time measurements with it. Secondly, I use a digital voltmeter (digital multimeter) for precise voltage or resistance tests. Finally, there is the home-made frequency counter from EN#40. That's all I need or use.

Your own setup may have slightly different requirements, but note that if you pick up a surplus scope for about \$50, you can get the other things that I have for little more than \$100 more. These things will not only get you going, but will keep you going for some time. Other things you can often borrow or obtain later. Note that you can probably use your synthesizer VCO's and other devices in your test program as well. Of course, you also have your ear, and thus an amplifier and speaker will serve as an important monitor of what is going on. I can't really work without hearing what is going on, and probably most musical engineers shouldn't even try to work without the speaker running.

\* \* \* \* \*

## ELECTRONOTES MUS. ENG. GROUP

APPLICATION NOTE NO. 14

213 DRYDEN ROAD

November 11, 1976

ITHACA, NY 14850

(607) - 273-8030

HOW TO ACTUALLY BUILD SOMETHINGPart 1: PARTS AND SUPPLIES

The first step in building any electrical circuit is to carefully study the schematic diagram and the circuit description to be sure you have a clear picture of how the circuit is supposed to function. The second thing is to be sure you have a clear idea of everything you will need to complete the project. These items fall into six general categories: electrical components (resistors, capacitors, IC's, transistors, diodes, etc.), control components (pots, switches), hardware (nuts, bolts, jacks), packaging (cabinets, panels, knobs), supplies (blank circuit board material, wire, solder, resist paint, etch solution, etc.) and tools (soldering iron, fingernail clippers, alligator clip for holding small parts during soldering, plastic dish for etching solution, drill, etc.).

## ELECTRICAL, CONTROL COMPONENTS, AND HARDWARE:

Unless you live in a large city or are only building a small project, you will probably find it easiest and cheapest to obtain the parts you need by mail order. To get addresses, start with ads in the back of magazines like Popular Electronics and Radio-Electronics. Also, get addresses from other builders. Electronotes also has a list of suggested suppliers. You will probably be able to obtain the semiconductors you need by mail in a week or two. You can also obtain resistors and capacitors in this way, but it is usually better to stock many of these items. Controls (pots and switches) are usually quite expensive as "new" items so it is well to watch for these on surplus lists and then buy some extras for your stock. Likewise, hardware items are not general surplus items so be sure to grab some when they do appear.

## SUPPLIES:

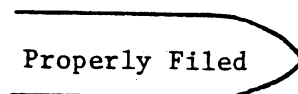
Items like circuit board material, wire, solder, paint, and etch solution are items which you are not likely to buy very often since you usually will be buying much more than you need for an individual project. Probably the best circuit board material is that which has fairly heavy copper, is coated on only one side, and which is already cut to some convenient size ( $4\frac{1}{2}$  by  $6\frac{1}{2}$  is common for example). It is not particularly easy to cut smaller pieces out of a very large board, so try to get smaller ones. It is convenient to get hookup wire that is of relatively small gauge (#24 or #26, and possibly stranded is most useful). Test the insulation. It should not melt under normal soldering iron heat, or at least should not melt easily. Yet the insulation should be soft enough so that you can easily strip it off with a thumbnail, as this will save much time and frustration looking for a wire stripper. Choosing a good grade of solder is very important. Solder is expensive, but the better grades are well worth the extra money as easier soldering means less time spent and fewer soldering errors due to awkward technique with poor grade solder. The solder should be rosin core, 60% tin 40% lead (or better still, 63% tin 37% lead true eutectic solder), and in a size of 1/16 to 1/32 inch diameter. Again we stress, do not buy cheap solder, as you will pay the price later.

You will need some sort of resist paint to paint up your circuit boards. It should be a paint of some sort - none of the "pen" type devices are good for more than about six square inches of board space, and when they dry out, you will uncover as much copper as you cover. You should rely on some kind of paint, and possibly press on tape if you happen to like the nice shapes and straight lines and don't mind the expense. You have a choice of paint, laquer or acrylic. Probably a lacquer paint is the best all around, but its biggest drawback is that the fumes of the solvent will irritate during an extended

painting session. Acrylic paint is water base and can be thinned with water and thus you don't have the solvent fumes, but the water base is less effective at coating the copper surface, especially if the surface is a little greasy or oily. In either case, the paint should be thinned down quite a bit before use. For laquer paint, add about 50% extra solvent (laquer solvent is available with laquer paint in most hardware stores). For acrylic paint, add about 25% extra water to start with, and possibly a drop or two of dish washing liquid as this will make the paint spread better. Acrylic paint is available at artist supply stores in tubes. Any color will work, but avoid browns, reddish browns, and dark yellows (which are similar in color to the etching solution and etching residues), and any color similar to the insulating material of the circuit board. The etch solution should be ferric chloride (the standard item) and you can get this in electronic supply stores. You can get this on your hands for a short time, but it should be washed off soon, and will stain just about anything, so be careful. Avoid "anhydrous" ferric chloride powder as mixing this with water is highly exothermic and fumes and dust can be most irritating.

#### TOOLS AND THE LIKE:

A pencil type soldering iron is best for IC work. The wattage of the iron should be quite low (the 27.5 watt size is good). It is extremely important that the iron and tip be kept in top shape. Of course, a new tip should be tinned with solder as soon as it gets hot. But you should also file and retin it if it becomes misshapen, which will generally happen in a matter of a couple of hours. Iron tips may last a little longer, but we generally suggest using the copper ones. A properly shaped tip is shown in Fig. 2 while a badly burned one is shown in Fig. 1. Often the burned one will have a concave tip and possibly a hooked needle like point that will make soldering stranded wire a nightmare. The properly formed convex tip will be easy to use. When the tip becomes misshapen, there is no need to cool the iron. Just twist out the tip with a pliers, allow to cool, file properly, replace, and tin immediately. Even if you don't have to refile the tip, it is a good idea to loosen the tip and retighten about every half hour to make sure it does not set up, and so that good thermal contact is maintained (many people think they need a hotter iron while all they need is better thermal contact between the heating element and the tip).



Two simple "tools" will be very useful. A fingernail clipper is probably the best of all possible cutters for small copper wire. It is much easier to use than a "diagonal cutting pliers" and is much cheaper (and often right in your pocket instead of somewhere on your bench). An ordinary alligator clip is the most useful device for holding parts while soldering (you hold a resistor in the clip and the clip in your hand and you don't burn your fingers!).

It's nice to have an electric drill handy for drilling holes when mounting boards and controls. A "rat tail" file is very handy for enlarging holes if you don't have the correct size drill. If you drill holes or file while some electrical components have already been installed, be sure to get the "chips" out. It is good to use an ordinary vacuum cleaner to catch the chips, and to clean up afterward.

The paint brush you use for etch resist paint should be found by trying several different ones. It is a good idea to get a fairly stiff one or to use one which has a little paint dried into it. Thus, it is sometimes desirable not to clean the brush after you use it.

Finally, it is necessary to have something to clean up the copper circuit board. This can be steel wool, or scouring powder, or something of that sort. All copper board you get will need to be cleaned both before you paint it and afterward. Be sure to shine it up well in both cases. If it is not cleaned well, it will be hard to get a good even coat of paint. If it is not again cleaned before soldering, the solder may not stick easily. A circular cleaning motion is best. This will also make it easy to mark the board with a pencil before painting.

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APPLICATION NOTE NO. 15

November 18, 1976

HOW TO ACTUALLY BUILD SOMETHING

Part 2: CIRCUIT BOARDS

After obtaining all the necessary parts and studying the circuit to be built, it is usually necessary to build a circuit board for the circuit. The circuit board serves as a mount for many of the components. Obtaining circuit boards presents a problem of a level of difficulty that varies with the individual builder. Some builders never prepare their own, but only buy them (and thus have a restricted number of circuits to work with). Others take days with photographic methods, and strive for professional looking boards. Some builders use a type of "instant circuit board" which is a prepared board set up for a general type of circuit. Still others simply paint and etch their own boards in a matter of hours. We tend to favor the last two of these methods because they get you going fastest even though the circuit may not be particularly efficient in the case of the "instant circuit board" and the results may not look as professional in the case of the hand painted board.

In our view, a fundamental mistake that many builders make when preparing a circuit board is that they look at commercial boards and decide that theirs should look as much as possible like the commercial ones. In particular, commercial boards have the components on one side, and the solder and traces on the other. This is fine, and often essential for production circuit boards (some commercial soldering techniques would prevent the placement of components on the foil side of the board). However, for the individual builder, use of both sides of the board in the commercial manner just serves to obscure half the circuit. In our view, it is absurd to go to the trouble of drilling a lot of small precisely placed holes so that you can hide the components from your view during construction and testing. Rather, we suggest placing all the components on the foil side of the board, and not drill any holes except those necessary for mounting the boards.

With the above paragraph in mind, we will describe a simple printed circuit board fabrication method that is fast and easy (and inexpensive). All that you will need to do this is blank circuit board material, some kind of etch resist paint, a paint brush, and some kind of cleaner such as steel wool or scouring powder. Next, one should adopt the view that the board he is about to produce will serve to mount the components, and to provide much of the interconnecting circuitry, but will not be the ultimate in efficiency for the particular circuit. Most likely, there will be a need for a few wire jumpers (which are the resistor lead clippings you would otherwise throw away). Also, there will probably be a need for a longer wire (or two, or three, ...) to do the job of a missing trace which you just could not find a path for. Of course, you can always find a better layout for any circuit if you keep trying, but if you do keep trying, you will never get the thing built.

So the first step is to insure proper mounting of the components. There will probably be no real need to worry about the board space for resistors, but only for IC's, large capacitors, and such items as trim pots. When this is determined, select a blank board large enough for the components in the circuit, and lay out the major components on the foil side of the board. Mark the positions of the leads of the IC's with a small pencil dot. Be a little careful about the layout of these major components. You should allow no more room than is necessary for the various sections, but then again you might as well use the whole board. Once you have the major components on the board and marked, make a quick check to see that everything is going to fill in well and adjust accordingly. Be sure to provide an area of the board for foil pads which will serve to receive wires from inputs and outputs, and from panel switches and controls.

Once the IC positions have been marked, you will know where the power supply lines must go. We suggest that these be penciled in using a straight edge. Later on, these straight lines will serve as references for other traces so that they can be made neater. With this done, the circuit board will look something like the one in Fig. 1. Actually, at this point it is just a copper clad board with pencil marks. You should note the following: (1) There is a ground path around the entire outside of the board, and at the corners this gets wider to provide mounting surfaces. (2) The dots represent the points where the IC pins will make contact with the copper surface. (3) At the left, a number of rectangular areas are provided to serve as in/out tabs. (4) The power supply lines have been penciled in for all the IC's.

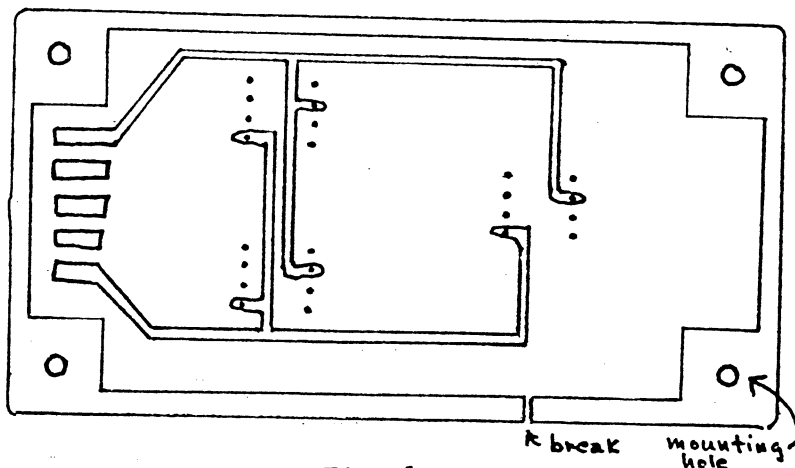


Fig. 1

The next step is to paint the board with either a lacquer paint (solvent base) or acrylic base paint (water thinner), and a thin artist brush. The paint should be quite thin, and the brush should be fairly stiff (just don't clean it out after each use). You of course paint the penciled in areas you have prepared. Then, you are pretty much on your own as far as completing the circuit is concerned. You just start painting in the necessary traces. Actually, you will soon find that many of the configurations you will need will become routine. The procedure is illustrated in Figures 2, 3, and 4. Fig. 2 shows an inverting amplifier circuit to be constructed. Fig. 3 shows the pattern the etch resist paint will take. In arriving at this pattern, you have in mind a mental picture of how the finished circuit will look (see Fig. 5).

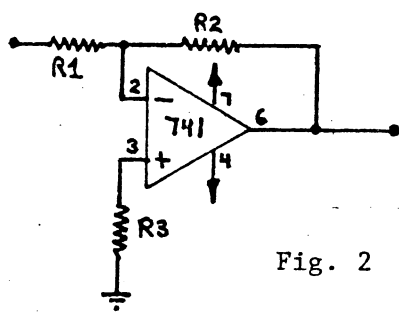


Fig. 2

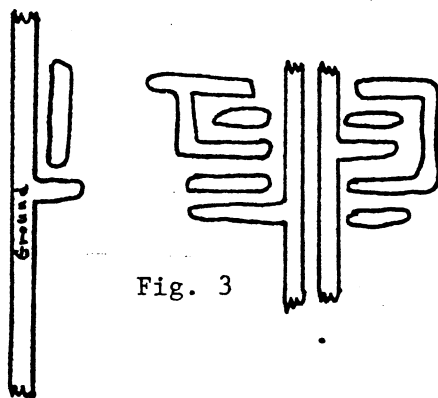


Fig. 3

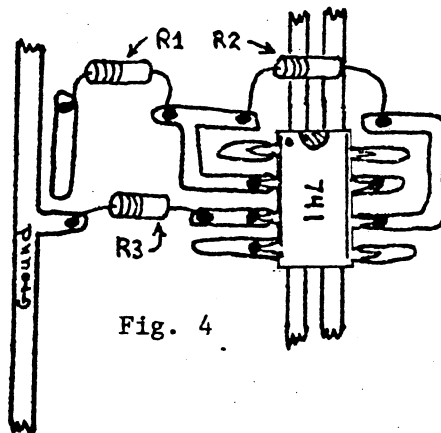


Fig. 4

Once all the paint is on and all the paint is dry, you etch the board in ferric chloride solution. It is usually possible to do this by floating the board on top of the solution. Touch the board to the liquid surface at a slight angle to prevent bubbles from being trapped underneath. Check the board after about 10 minutes, and replace if necessary. Once etching is complete, rinse the board with water and remove the paint with solvent or steel wool. Then thoroughly clean and shine the board with steel wool or scouring powder. Rinse with water and dry.

You can begin soldering parts to the board immediately if you wish, or you can take a little time to tin the board first. To tin the board, you just coat all the copper with a thin coat of solder. This will repair any small cracks in the copper, and will serve to make part mounting easier. If you wish, you can then invert the board and remove excess solder by letting it run down the soldering iron tip. This leaves a very thin coat which can be polished with steel wool to look like a tinned board.



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APPLICATION NOTE NO. 16

November 25, 1976

HOW TO ACTUALLY BUILD SOMETHING

Part 3: SOLDERING IN PARTS

Whether or not the actual soldering in of parts is easy or hard depends to a large degree in how successful you were at preparing a circuit board. If we assume that you had a good mental picture of the way the parts were to fit in when you painted the board, this part of the job is just a realization of the mental picture. In this note we will be talking about the best techniques for soldering, some of the things to do when things don't go exactly as planned, and a few of the nice things you can do to make testing easier.

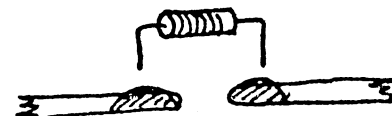
We talked in an earlier note about the importance of good solder and an iron in good condition. These are the first of the important "rules" of good soldering. The third rule is to not use too much solder. If you get big blobs of solder on a circuit board where IC pins are close together and circuit board traces are also close, you are going to have problems. Not that you are likely to miss seeing a blob that would short the circuitry, but they are very hard to get rid of. The best thing is to avoid them by using only small amounts of solder, but if you do get one, you may be able to get rid of it by holding the board upside down and trying to get the excess solder to run down the iron tip. The fourth rule of good soldering is to not use too much heat. It is not likely that excess heat will damage any components, but the point is that you should not have to use it. You should only have to touch the soldering iron to the connection for a period of a second or two. By the way, there used to be instructions for soldering that said you use the iron to heat the metal to be soldered, and then touch the solder to this metal (instead of touching the solder to the iron). This never really was too successful, and it took too much time. Thus, even when these stories were going around (as in Heathkit instructions), no one really did this. It is much more realistic to touch the iron to the metal, and push the solder in between the iron and the metal, letting the solder flow. If you then get the soldering iron out fairly fast (before all the rosin flux burns off), you will get a smooth connection. The fifth rule of soldering is that things are easier to solder if they are tinned first. That's one reason why we suggest tinning the board first. Now, you probably know that the leads of resistors and capacitors are already tinned, so you might think that there is nothing more to do with these. Actually, a freshly tinned layer will work much better than just relying on the original tinning. The same goes for IC pins which will be much easier to solder if you retin them before attempting to mount them.

At risk of boring the reader, we will go into a detailed discussion of how a single resistor may be mounted. Assume that you have a position for the resistor and an uncut resistor as shown in Fig. 1. The first step is to tin the end points of the circuit board traces with a little blob of solder. Next you bend over the resistor leads and cut them at a point about 1/4 inch beyond the bend. The spacing between these cut ends should be the same as that between the mounting points (Fig. 2). Next, grasp the body of the resistor with an alligator clip. Hold this in one hand and the soldering iron in the other. You should also have a coil of solder with a free end sticking out. Touch the cut end of the resistor, the soldering iron tip, and the solder all together at the same time, and then pull them apart. Done



Fig. 1 Mounting position and uncut resistor

Fig. 2 Tinned mounting position and cut resistor lead.



properly, this will result in a deposit of solder on the lead end that ranges from a heavy coat to a small blob. Then you place the resistor (still holding with the alligator clip) above the mounting position and press down slightly. Touch the soldering iron tip to the junction for a period of about one second, and then remove. Hold the resistor in place until the solder hardens. Now, the other end of the resistor is held in approximately the right position by itself. If there is enough solder at the junction, you just heat the junction and remove the iron. If not, bring up a little more solder. Then, after this second junction hardens, if there is any doubt at all about the first one (you may have moved your hand slightly while waiting for it to harden), you should reheat the first one, perhaps adding another bit of new solder. Be sure you don't have both ends in liquid solder at the same time or the resistor falls over. The entire mounting may take as little as 20 seconds. The final soldering procedure is illustrated in Fig. 3.

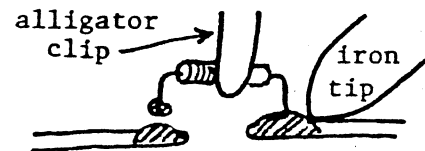


Fig. 3a First Connection

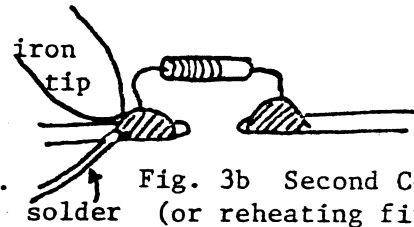
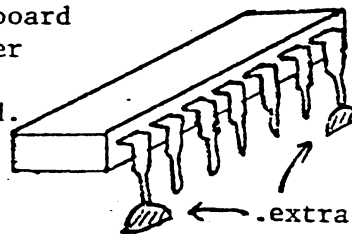


Fig. 3b Second Connection  
solder (or reheating first one)

IC's and other components are mounted in much the same way. You first tin all the metal parts, then get the part "tacked soldered" in place, and finally reheat and add solder to any additional connections or poor connections. Eight-pin "Mini-DIP" IC's can generally be mounted flush with the board. For 14 and 16 pin DIP packages, it may be a good idea to mount them slightly above the board surface by first adding a larger blob of solder to the corner mounting positions, and not forcing the IC to the board as solder is added. This makes it easier to remove should this become necessary. This elevated mount is shown in Fig. 4.



Elevated Mount for  
14 pin or 16 pin DIP

Fig. 4

So far we have assumed things have gone as planned. They won't always. If you have an improper trace, you can usually break it with a pointy knife blade. If you are missing a trace, use some wire. If you forget to leave mounting tabs, you may be able to sneak a component in "upright" (Fig. 5), or may even have to establish a "floating" tie point (Fig. 6). It is even possible to "piggy-back" an IC to another (using common supply pins) as shown in Fig. 7. This is usually easy to do, and can be made quite neat and mechanically sound. There are numerous other ways to improvise and get the job done. If most of the circuit is correctly prepared, you should be able to get out of a few tight spots.

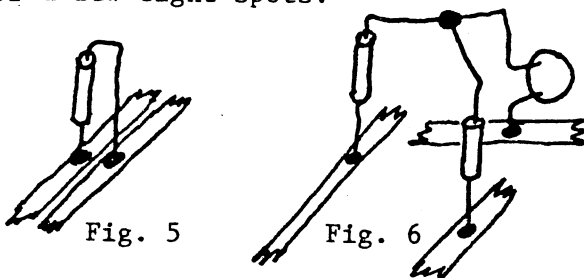


Fig. 5

Fig. 6

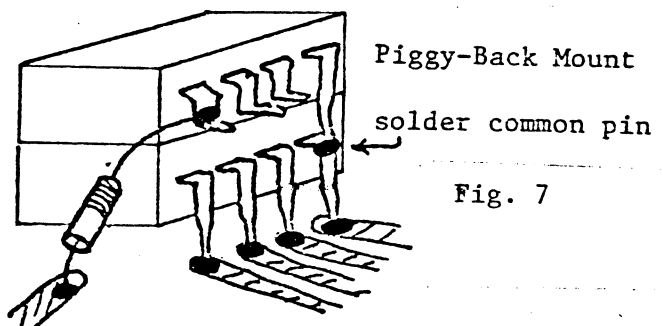


Fig. 7

It is a good idea to give some thought to the testing phase that will follow. At certain critical points in the circuit, a resistor lead mounted straight up may be a useful point to clip on a test lead (Fig. 8). You can remove these later or just leave them in. At certain points, it may be best to leave a jumper which can be removed to isolate major sections of the circuitry during testing. These jumpers can be looped up so that a test clip can be attached to them as well.

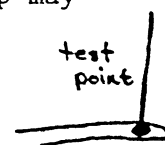


Fig. 8

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APPLICATION NOTE NO. 17

December 4, 1976

HOW TO ACTUALLY BUILD SOMETHING

Part 4: PACKAGING

By the time you get this far, you probably have a circuit board built and tested and are looking for something to install it in. Or, you may be waiting to test it because you need to connect it to certain panel controls that will be mounted as part of the package, and not on the circuit board. If at all possible, test the board outside. You can often do this by temporarily soldering in a fixed resistor or two as a substitute for a panel pot. It will help a lot if you know the board going in does work, or at least is not totally unknown as far as its operation is concerned.

So what is packaging? It is just putting the circuit board inside some sort of box or enclosure to protect it from damage and make it easy to use. You may be proud of your circuit, but the advantages of putting it in a box are obvious. Or, you may want the box to hide a circuit that is electrically perfect but otherwise not what you would call a work of art. You will probably find that enclosures are not as easy to come by as are the components inside, and you may well end up spending more for the box than for the rest of the circuit. Naturally you will want an enclosure that is big enough to contain the circuit board, the power supply, and whatever else must go inside. It should also provide the necessary mechanical support and shielding. Other than that, you will only have to consider the ease with which you can finish the project (and of course the price and availability). You could easily spend \$10 or more for an enclosure, or you may find a thrown away box that will work.

Let's assume that your board is of the general form shown in AN-15, Fig. 1 (a board with mounting areas and holes in the corners. Since you have all the parts on the foil side, nothing on the back, and have drilled no holes other than the mounting holes, all you have to do is drill matching mounting holes in the enclosure and mount the board with four nuts and bolts. If your enclosure is metal and serves as ground, you have grounded your board. Next, you will run up the wires from the power supply to the board. It is often a good idea to add a bypass capacitor to these supply lines at the point where they come on the board. This helps to keep noise from entering the board through these lines, and noise generated on this board off the lines. It is often the case that individual circuit boards may work just fine when tested separately, but fail when mounted in an enclosure with other boards on a common supply. These bypass capacitors (which should be 5 to 10 mfd electrolytic for example) will help to assure you may not have this problem.

Now, for a typical device, you probably have a front panel that has on it pots, switches, jacks, and possibly pilot lights and other indicators. If you have a power supply that runs on the AC line, you should do your best to place the power supply switch out of the way (perhaps under a chassis) so that you won't accidentally come in contact with high voltage. Once high voltage lines are soldered, make sure the power supply is functioning properly, and then mix up some epoxy glue and coat the exposed terminals. Let this glue dry, and then cover the terminals with electrical tape as well. There is no sense taking chances, and builders these days have a tendency to forget about high voltage because they are usually using only low ones. If a wire end strays to the AC line, even if you don't get hurt yourself, you may have components that will object. If you take these steps, you have a safe supply setup, and your board is powered properly much as it would be on the test bench. You can then install the additional control wires and test them one at a time as you go.

Once the board is installed, and all control pots, switches, and jacks are installed, you will be faced with the obvious problem of connecting the controls to the boards, and wire of some sort ~~are~~ the obvious solution. It is well to do at least a little planning at this point. One of the major rules here is that there is generally no reason to run a wire from a tab on the board to the panel more than once. That is, you run to the panel once, and if you need the same line again, you jump from one panel location to another. Probably stranded wire is best for connections from the board to the panel as it is flexible. On the panel however, solid wire is best for jumping from one point to another as it stays in place better. A couple of examples will illustrate the above points. Fig. 1 shows a pot acting as a voltage divider providing from +15 to ground. It is often the case that the panel is grounded, and thus, so is the case of the pot. We simply bend terminal three over and solder it to the case, leaving us with only two wires to run to the board. In Fig. 2, we show two voltage divider pots between +15 and -15. We run the +15 and -15 to the panel at only two points (one for + and one for -) and jump from there to the various pots. Thus, for each additional pot we connect up, there is only one additional wire running back to the board.

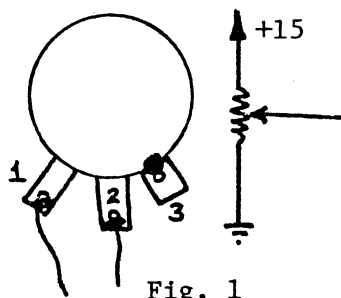


Fig. 1

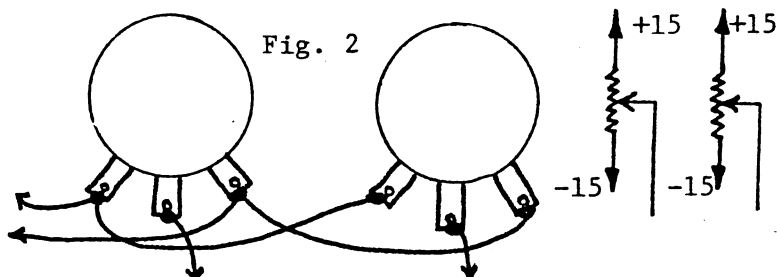


Fig. 2

As the wires come off the board, it is usually a good idea to route them through certain paths so that they don't jump all around. There are probably plenty of places where you can do this. Fig. 3 shows three possible paths. At A, the wire runs along and under an IC which is mounted on the foil side of the board with its leads flush. At B, we have used any handy resistor and run the wire underneath. At C, we show how a loop of wire can be used as a threading point. It is just a small loop of solid insulated wire (or just an uninsulated resistor lead clipping) which is looped over and soldered in two places to the same conducting path. This may be a jumper you are already using. Short wires soldered only at one end are also useful for collecting loose wires and lacing it around them holding them in place. Threading points on the panel will not be quite so easy in general. You can loop a wire and connect both ends to a pot terminal for example. Or, you may have a panel jumper wire which is too long as is. Give it a little twist as shown in Fig. 4, and you tighten this up and provide a threading point as well.

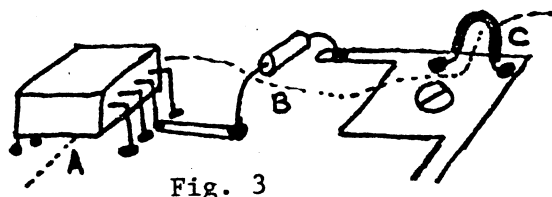


Fig. 3

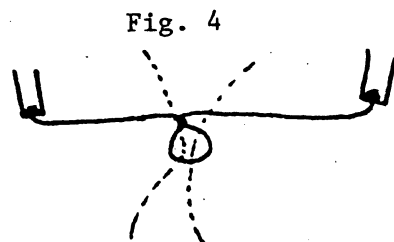


Fig. 4

You should also consider the possibility that you can mount components on the panel. Fig. 5 shows a common situation where a rotary switch selects one of seven possible resistors. If the resistors are on the board, you need eight wires between the panel and the board. If you mount them on the switch, you have all the resistors coming to a common floating point which is just a solder blob, and only two wires between the panel and the board (x and y).

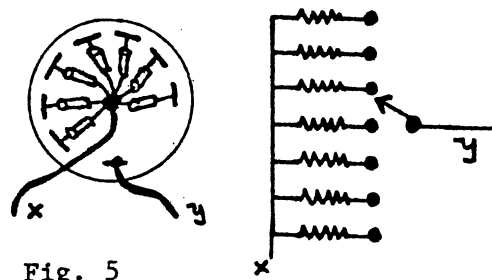


Fig. 5

There are many other tricks which you will surely devise. Also consider the use of twisted pairs or ribbon cable. To make a twisted pair, just stretch out two wires across the room securing the far end. Put the near ends in an electric drill chuck and pull the trigger - instant twisted pairs!

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APPLICATION NOTE NO. 18

December 11, 1976

HOW TO ACTUALLY BUILD SOMETHING

Part 5: MISCELLANEOUS HINTS

In preparing parts 1 to 4 of this series (AN-14 to AN-17), we had a specific phase of construction in mind for each note. Here we want to clean up a few loose ends which are also important. These are presented in no particular order.

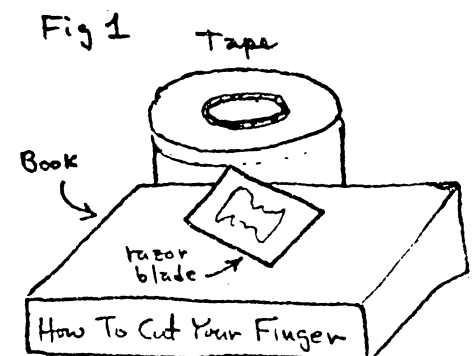
UNUSED (N.C.) IC PINS: When a pin of an IC is a real NC (no connection), there is of course no reason to solder it down, as this will just make it more difficult to remove. On the other hand, it is a good idea to leave a blank tab on the board for it. Why? It serves as a reference to the eye when installing IC's on the board. A typical op-amp may use only 5 of eight pins for example. It is much easier to place this if there are two parallel rows of eight pins than it is if there are only three on one side and two on the other. This is especially important if you have IC's in a row with spaces in between as otherwise a blank space may look like the space between IC's. When installing round IC packages however, it may be a good idea to solder down all NC pins as well as this prevents these wires from flying around, and they are not hard to desolder.

MARKINGS ON PANELS: When you have to put lettering on panels, rub off type letters are a good choice. You will have to put these letters on before mounting any controls on the panel however. Be sure the panel is clean of any grease or oil or you will have a very hard time with these. Once on, several light coats of clear lacquer will prevent them from rubbing off. A coating of clear nail polish is also good for protecting these letters, but brush this on very lightly or you may dissolve the letter.

HOLDING ETCHING SOLUTION: The best thing I know of for etching solution is a plastic freezing dish with press-on cover. This allows you to keep the solution covered when not in use. The dish should be big enough for your biggest board, and not too deep.

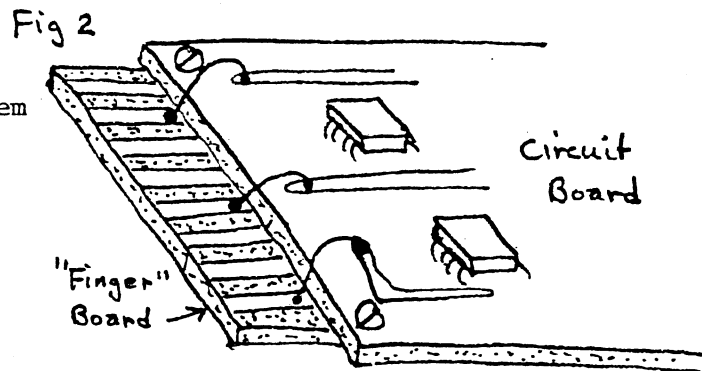
CUTTING CIRCUIT BOARD: If at all possible, avoid cutting circuit board. When it is necessary, it should be sawed and not sheared unless you have a shop type sheet metal shear available. An ordinary "tinner's shears" will not work as it will leave a ragged edge, will warp the board, and leave cracks in the copper. A hack saw will work as well as anything if you don't have a band saw available. If this sawing leaves an uneven edge which you don't like, set a sheet of sandpaper on a flat surface and rub the edge on the sandpaper. This results in a very nice even edge.

PRESS-ON TAPE FOR CIRCUIT BOARDS: For making straight edges, press on resist tape usually beats painting. This is expensive. Electricians plastic tape works just as well but is too wide as it comes. A good way to cut this to a proper width is to start with a wide roll such as 3/4 inch and place this on a flat surface. Bring up a thin board, writing pad, book or whatever such that the top surface of the book is flush with the surface at which you would cut the roll. Then place a razor blade flat on this surface, and hold it slightly over the edge. Rotate the roll of tape against the blade and you will be able to make a fairly even cut. Don't try to cut the whole roll at one time - cut some and use it and then go back for more. When you get into the top slice a ways, get a thinner book and work on the next slice a while, and so on. This procedure is illustrated by Fig. 1.

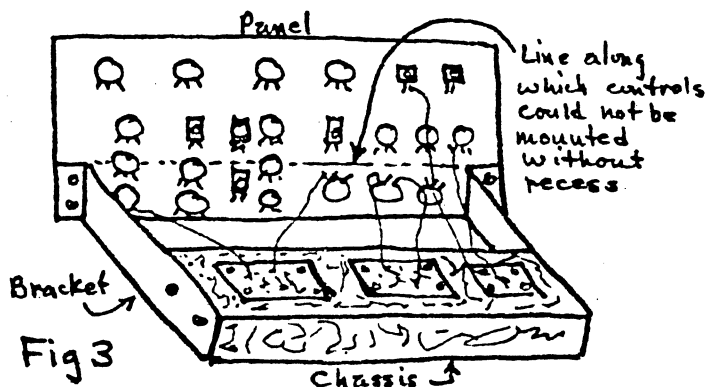


MORE ON PREPARING CIRCUIT BOARDS: We mentioned that it is often useful to tin the entire circuit board after preparation. In general you just use it as is after it is tinned, but if you wish you can make the board look as though it were solder plated. To do this, you must have removed the excess solder from the tinning process by holding the board upside down and letting it run down the soldering iron. This leaves a shiny and possibly "grainy" surface appearance. The board can then be washed with alcohol to remove excess rosin flux. Next, you use steel wool to smoothen up the board. This would also remove any excess rosin flux, but the flux tends to clog up the steel wool so it is best to get it off first. After getting the grain off the surface, polish the board with scouring powder, and you get a nice looking "plated" board for your efforts.

FINGERS FOR EDGE CONNECTORS: If you are mounting your PC boards in edge connectors, there must be a matching set of fingers on the boards. These can be painted on, but it is difficult to measure properly each time you need these. One easy way is to just insert the board in the connector once or twice and this will leave a mark on the board showing where the connector contacts will meet the fingers on the board. You can then paint them in. Another thing that is useful is to make a whole board of fingers by using press on tape. This is especially useful if you have some type of instant breadboard which you prefer but which does not have fingers. You can then cut off a slice of the finger board and bolt it to your board. The general idea is shown in Fig. 2.



OPTIMUM USE OF PANEL SPACE: With electronic music equipment, and with other types of equipment as well, panel space is at a premium because there are so many controls we need to have available. If we have a chassis mounted directly to the back of a panel, there is a line on the front of the panel along which we are not able to mount any controls because the chassis top is directly behind it. For this reason, it is often desirable to recess the chassis back from the panel so that every inch of available space can be used to maximum advantage. In doing things this way, we can just lay out the panel from the front and not have to think much about what is to go behind. We also avoid problems of drilling mounting holes through both the panel and the chassis and having to worry about whether or not they will line up. The chassis may be recessed by a set of brackets which you can buy as side braces for a flush chassis, or you can probably make them yourself. The basic idea should be clear from looking at Fig. 3.



GROUNDING AND BYPASS: If you are making your own circuits, you do not have a tried and tested layout to work with. If you are building high level signal handling gear such as synthesizer modules, you should have few layout problems. Digital circuits may be a little trickier, but again probably no real problem. The major problems with layout will occur with low level audio circuitry, and with radio-frequency devices. Many problems are avoided by proper grounding and bypass techniques. The major rule with grounds is to make them heavy, direct, and all to the same point. That is, all grounding paths radiate out as a star from a central point. Often, low current grounds should take different paths than those carrying large currents. In many cases, any old ground will do, but it is well to play safe. You can always make grounds heavier with an extra coat of solder, or by soldering down an extra wire to the trace. Supply lines should be bypassed at critical points. Often these points are located by noisy waveforms. Always bypass lines in digital circuits, and bypass the supplies to op-amps that are handling low level signals as close to the chip as possible.

## PREFERRED CIRCUIT COLLECTION

Power Supplies  
Controller Interfaces  
Voltage-Controlled Oscillators  
Manually Controlled Oscillators  
Voltage-Controlled Amplifiers  
Envelope Followers  
Voltage-Controlled Filters  
Timbre Modulators  
Envelope Generators  
Balanced Modulators  
Frequency Shifters  
Sample-and-Hold Units  
Slewing Circuits  
Noise Sources  
Analog Delay Lines  
Animators

#### A Note on Documentation of These Circuits:

Our main aim in the original "Preferred Circuits Collection" was to give all the necessary circuit diagrams, and whatever text material came along in the process. In this present collection, we have removed a few circuits, and expanded the text material a good deal. In most cases, this is accomplished by just including the original material in entirety, text and diagrams. In other cases however, this is awkward because the text relevant to construction is intermixed with more general theoretical material, and in such cases, brief notes written specifically for this collection have been added.



## POWER SUPPLIES

Power Supplies: The standard power supply for most of our designs is one that supplies +15 volts and -15 volts, referenced to ground. This is called a bipolar  $\pm 15$  supply. Some modules will also use a +5 volt supply. If you build a supply rated at one amp from each voltage, this is in general all you will need as it will drive your very first module and will continue to drive your system as it grows. A number of other supplies are suggested as well, which may be suitable in smaller "stand alone" units. The supplies and regulators listed in this collection are:

| <u>OPTION</u> | <u>SOURCE</u> | <u>DESIGNER</u> | <u>APPROX COST*</u> | <u>NOTES</u>           |
|---------------|---------------|-----------------|---------------------|------------------------|
| PS-1          | AN-1          | Hutchins        | \$6.50              | +5 volts, 1 amp        |
| PS-2          | AN-2          | Hutchins        | \$11                | $\pm 15$ volts, 1 amp  |
| PS-3          | AN-98         | Hutchins        | \$7                 | $\pm 15$ volts, 100 ma |
| PS-4          | AN-136        | Hutchins        | \$6.50              | +15 50ma, -14 10ma     |
| PS-5-Acc      | MEH-5j        | Rossum          | ----                | Crowbar circuit        |

\*includes transformer, but not cord, fuse, switch, or cabinet

The supply, PS-2 will be your main supply, while you will use PS-1 as well if you need +5 volts. PS-3 is a lighter duty supply, and is useful for stand-alone units. Similar light duty units can be made with the type 4195 and 1568 regulators. PS-4 is a simple, very light duty supply useful where a very small current at -15 (-14 actually, but close enough) is required. PS-5-Acc is an accessory for a power supply, a "crowbar" or overvoltage protection circuit. If the voltage tries to go much above +15 volts, for whatever reason, the crowbar will short the line to ground, protecting the attached circuitry.

It is well to take some care with the construction of the power supply because this is the one unit where dangerous line voltages are present. Use only grounded 3-wire line cords, and a fuse of from 1/4 to 1/2 amp should be used. Once constructed, all exposed high voltage connections should be taped, and then seal up the unit's cabinet. Pilot lights, while not shown, are also useful of course.



## ELECTRONOTES MUS. ENG. GROUP

APPLICATION NOTE NO. 1

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August 4, 1976

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FIVE-VOLT ONE-AMP POWER SUPPLIES FOR TTL

The following notes give step-by-step instructions for making a one amp power supply based on the LM309K +5 volt three-terminal voltage regulator, or you can use the LM340 +5 regulator.

(1) The first step is the selection of a transformer. For this circuit, a transformer rated at a voltage from about 8 to 16 volts at one amp can be used. A standard 12.6 volt "filament" transformer can be used (Ignore the center tap wire if there is one - cut off and tape end, or connect to a blank terminal somewhere.). The higher the voltage of the transformer, the smaller the filter capacitor can be. Thus higher voltage transformers (up to 16 volts) are suggested where portable supplies are needed and lower voltages (8 volts) where stationary supplies are used since the larger filter capacitor can be tolerated in such cases, and less ripple and heat dissipation result in this setup. The AC line should be connected to the primary of the transformer and the proper output voltage verified (using for example, a simple multimeter).

(2) The next step is to convert the AC voltage from the secondary to DC. This forms an overvoltage, unregulated DC supply. (Note that the primary leads to a transformer are usually black while secondary leads are usually colored red, green, or yellow and a center tap is usually of mixed coloring). The full-wave "bridge" rectifier structure formed from four diodes is suggested. The diodes should be rated at one amp or higher, and at 100 volts or higher. (1N4002, 1N4003, 1N4004, 1N4005, 1N4820, etc., or a DIP bridge are good choices). See step two of the figure. After installing the bridge, the voltage output will read on a DC meter scale at a voltage in the neighborhood of the transformer secondary voltage.

(3) Next choose the filter capacitor. The formulas to do this can be found in Chapter 5J of the Musical Engineer's Handbook. For this application, just use the table below:

| <u>Transformer Voltage</u> | <u>Nominal Capacitor</u> | <u>Suggested Value = C1</u> |
|----------------------------|--------------------------|-----------------------------|
| 8                          | 2600 mfd                 | 2200 to 4000 mfd 16v        |
| 10                         | 1400 mfd                 | 2000 mfd 25v                |
| 12.6                       | 820 mfd                  | 1000 mfd, 25v (35v OK)      |
| 14                         | 720 mfd                  | 1000 mfd, 25v (35v OK)      |
| 16                         | 580 mfd                  | 1000 mfd, 35v               |

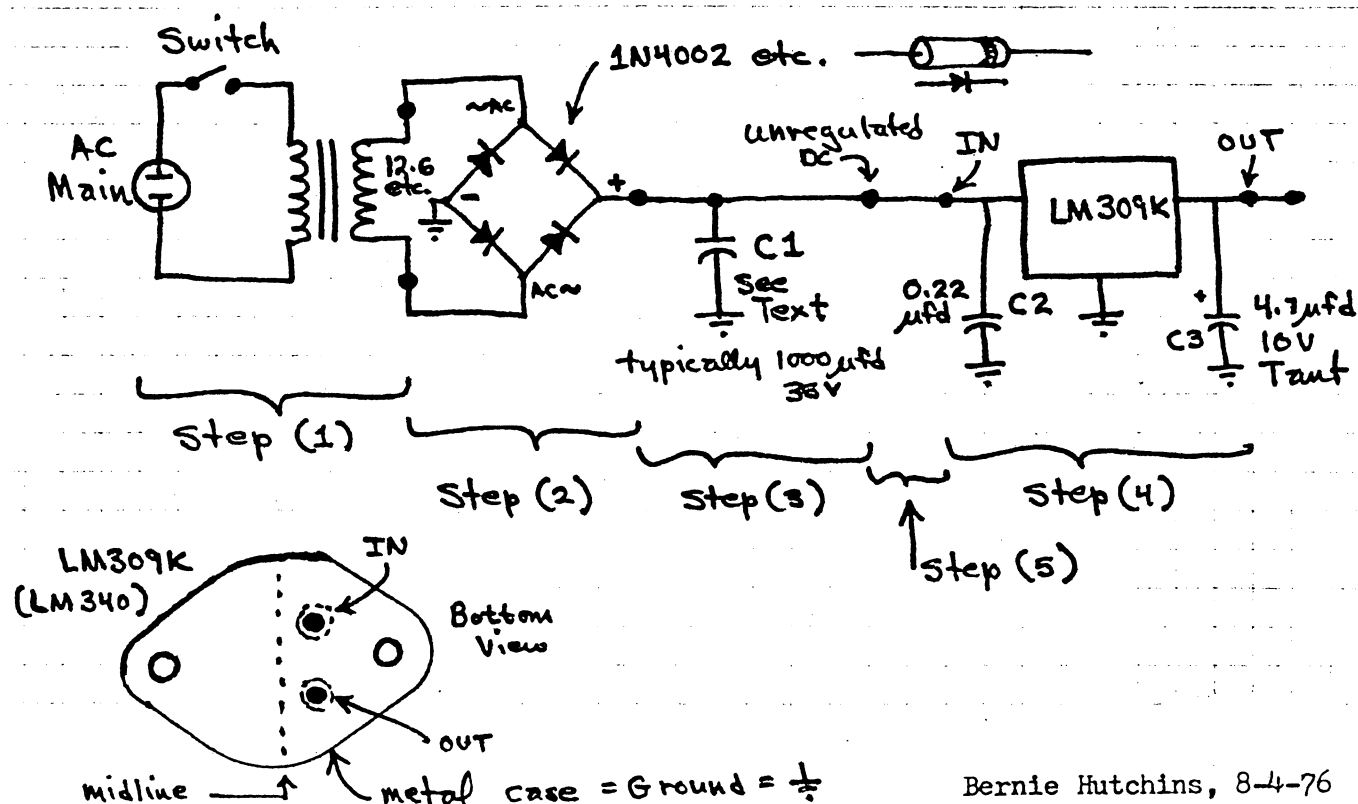
The values above are somewhat conservative, and are for a full one amp output current. The capacitor value may be decreased in proportion to the actual output current. For example, if your circuit requires only 300 ma (0.3 amps), you can multiply the suggested values by 0.3. The power requirements for TTL can be obtained from TTL data books. For portable devices, the required current is often well determined. For lab type and experimenter supplies, the current varies with the device being tested, and it is probably best to use a capacitor so that the full one amp may be drawn. Connect up the capacitor making sure the polarity is correct (capacitor has either the + or - end marked). Turn on the power and measure the DC voltage. It should be about 1.4 times the transformer secondary voltage. This is because the capacitor holds the peak voltage of the AC cycle which is  $1.414 \dots$  times the rated (RMS) value of the AC secondary. After verifying, turn the power off, and then short the capacitor. This will result in a moderate spark if you short it with a piece of wire, screwdriver, etc. You should discharge it now so that you don't surprise yourself when installing the regulator. The spark won't hurt you, but if you don't expect it, you may hurt yourself when you jump away. If you don't like sparks, short the capacitor with a 1000 ohm resistor, holding this in place about 10 seconds.

(4) Next install the regulator. This is the LM309K (or LM340 +5) which is a three terminal device. It has an input (from the unregulated DC supply you just made), an output (the +5 volts you want), and a ground. The regulator is particularly convenient since the ground "terminal" is the metal case. This means that you can mount the regulator in a metal chassis (which is usually ground) and the metal chassis will serve to give the necessary heat sinking to the regulator. Whether or not you need a heat sink depends on the amount of power being dissipated. The power is the product of the current being drawn out, and the voltage across the regulator. The voltage across the regulator is the unregulated DC input minus the +5 volt output. As a rule of thumb, if the current is 100 ma or less, no heat sink is required (the regulator case is enough). If the current is 500 ma or more, a heat sink should be used (that is, some extra heat sinking, metal chassis, metal plate, commercial power transistor type heat sink, etc.) If in doubt, calculate the power. If it's above 2 watts, use some heat sinking. In operation, the regulator should not be too hot to touch. We also suggest that the input and output capacitors C2 and C3 be used although they are not always necessary. Install these right on the regulator pins.

(5) Connect the unregulated DC to the input of the regulator and turn the supply on. The output of the regulator should be near +5 volts (from 4.8 to 5.2). The input to the regulator should be the same voltage it was before it was connected to the regulator (1.4 times the AC secondary). Next, connect up your load. The output voltage should remain at ~+5, or change by no more than a few tens of millivolts. The input voltage to the regulator will drop when the load is connected. The reading when read on ordinary meters should not drop below the midpoint voltage between the unregulated DC input (1.4 times the transformer secondary) and +7. For example, if you use a 12.6 volt transformer, the unregulated DC is about +17. Under full load, the unregulated input to the regulator should not read below +12. Also check the regulator for overheating after it has been on for 10 minutes or longer.

# DIAGRAMS:

PS-1



Bernie Hutchins, 8-4-76

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APPLICATION NOTE NO. 2

August 11, 1976

BIPOLAR 15 VOLT SUPPLIES FOR OP-AMPS

The following notes give step-by-step instructions for making a bipolar supply, +15 and -15 volts, for op-amp circuits. The circuit is based on the LM340 and LM320 type regulators. There will be three leads coming out of the supply, +15, -15, and ground. The supply will deliver one amp current from each voltage. You could make a bipolar supply by connecting the + terminal of one supply to the - terminal of another, and call this connection point ground. This requires two supplies, and this means two transformers, or at least two separate secondary windings on the transformer. Here we will use the center tapped type of transformer, a bridge rectifier, and a positive or negative voltage regulator depending on which half the supply we are working on.

(1) The first step is the selection of a suitable transformer. The transformer should be rated at one amp and have a voltage from 28 to 48 volts, with the values between 30 and 40 being preferred. The transformer must have a center tap on the secondary for this application. Note that standard "filament" transformers are not suitable for this application since they come in only 12.6 and 24 volt sizes and these are outside the usable range, even when used in series combinations. In an emergency, three 12.6 volt transformers in series can be used for a 37.8 volt center tapped transformer with the center tap of the center transformer being the center tap of the combination. The AC line is connected to the primary of the transformer which is usually two black wires. The secondary wires are usually colored, and the center tap is usually of mixed coloring.

(2) The next step is to convert the AC voltage from the secondary to DC. This is done using a bridge rectifier formed from four diodes. The diodes should be rated at one amp or higher, and 100 volts or higher. (1N4002, 1N4003, 1N4004, 1N4005, 1N4820, etc. are good choices for diodes.) The diodes are arranged as in step two of the figure. Note that the center tap of the transformer is grounded. Check the setup using a DC scale on a multimeter. Connect the - terminal of the meter to ground, and the + to the + side of the bridge. The DC reading should be about half the secondary voltage. Repeat connecting the + terminal of the meter to ground, and the - terminal to the - side of the bridge.

(3) Next choose the filter capacitor. You will need two of these. For a full one amp supply, the capacitor should be matched to the voltage of the transformer secondary as follows:

| <u>Transformer Voltage</u> | <u>Nominal Capacitor</u> | <u>Suggested Value</u> |
|----------------------------|--------------------------|------------------------|
| 15 volts (30 CT)           | 2080 mfd                 | 2200 mfd 35v           |
| 17 volts (34 CT)           | 1225 mfd                 | 1000 mfd 35v           |
| 18 volts (36 CT)           | 1016 mfd                 | 1000 mfd 35v           |
| 20 volts (40 CT)           | 758 mfd                  | 1000 mfd 35v           |
| 22 volts (44 CT)           | 604 mfd                  | 1000 mfd 35v           |

Alternatives:

|                    |          |              |
|--------------------|----------|--------------|
| 14 volts (28 CT)*  | 3205 mfd | 4000 mfd 25v |
| 24 volts (48 CT)** | 502 mfd  | 470 mfd 50v  |

\*suggested for fixed supplies - has lowest heat dissipation

\*\*suggested only if first line, non-surplus, regulators are used.

These capacitors are installed as shown in step three of the diagram. Make sure the + and - terminals are connected correctly. Turn on the power and measure the voltages on the capacitors. The voltage should be 0.7 times the transformer secondary. (That's 1.4 times the center tapped voltage - right?). Typical values are shown on the diagram. After verifying the correct voltages, shut off the power and discharge

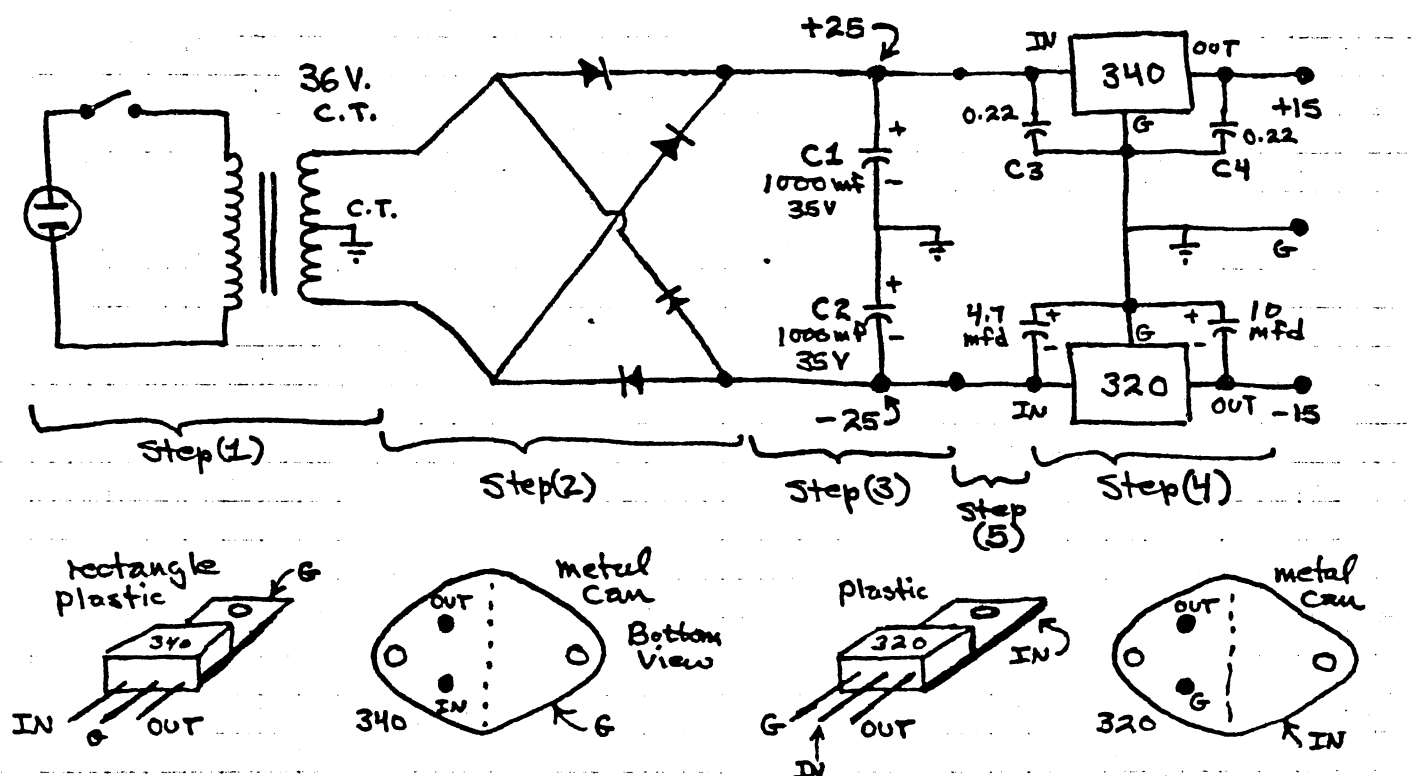
both filter capacitors by shorting them with a screwdriver or piece of wire. If you don't like sparks, discharge them with a 1000 ohm resistor held in place for about 10 seconds.

(4) Next install the regulators. The LM340 is the +15 regulator and may come in the metal can with oblong base, or in the rectangular plastic package. This is easy to install in either case, as the large metal surface is the ground terminal. This makes it easy to heat sink. Some sort of heat sinking is necessary unless only about 100 ma or less is to be drawn. See AN-1, step 4 for more information on heat sinking. Install stabilizing capacitors C3 and C4 as shown. These should be mounted right on the regulator terminals. The LM320 is a little trickier to use since it is more difficult to heat sink, and because it requires larger stabilizing capacitors. The large metal surfaces of the LM320 are the input terminal, not ground. Thus, if heat sinking must be used, the heat sink must be insulated from ground. Alternatively, you can insulate the metal surface from the chassis with a mica washer and some insulated mounts. For currents from -15 of 100 ma or less, the LM320 rectangular plastic package is nice since it requires only one bolt to mount, or can be soldered in "standing up." Be sure to include the stabilizing capacitors on the LM320, and keep their leads as short as possible. Solder them right to the regulator pins.

(5) Connect the unregulated DC voltages to the inputs of the regulators. Turn on the power and measure the output voltages. Both the +15 and the -15 should be within 0.6 volts of their nominal values. The voltages at the inputs to the regulators should be the same as they were when they were not attached to the regulators. Next connect up the load and the output voltages should not change very much (less than 3%). The input voltage may drop with the load connected, but should not drop below the midpoint of its original value and +17 volts. Check the regulator for overheating after it has been on a while. It is a good idea to load the LM320 with about 50% more than the expected current drain to see if it does anything funny. If there is no problem, you can go back to the original load with some assurance that things are working well. The LM340 will in general be less of a problem than the LM320.

# DIAGRAMS:

PS-2



1 PHEASANT LANE

August 11, 1978

ITHACA, NY 14850

IC REGULATORS FOR SMALL BENCH SUPPLIES

(607)-273-8030

In this note, we will be discussing some IC regulator circuits suitable for use in small bench supplies or in the "transferable" power supply discussed in AN-97. We will begin with a discussion of a more general problem in supply regulation - the problem of keeping the heat generated in the IC down to acceptable levels.

A typical regulator setup is shown in Fig. 1. The regulator is supplied by an input voltage from an unregulated supply. The unregulated supply consists of a transformer, diodes, and the filter capacitor  $C$  as shown. This unregulated voltage can be thought of as always coming from the capacitor, and this capacitor is charged each time the AC cycle goes to maximum value. Between extremes in the AC cycle, current flows from the capacitor and hence the voltage on the capacitor drops. There is thus considerable "ripple" on the input, as shown in Fig. 2. It is the job of the regulator to take this input and "slice" off the excess voltage, ripple and all. As should seem reasonable, this "slicing" action will cause some heating of the regulator. In fact, either Fig. 1 or Fig. 2 will show you that there is a voltage of  $V_{in} - V_{out}$  across the regulator, and the current through the regulator is the output current  $I$ . Thus the power dissipated across the regulator is  $(V_{in} - V_{out})I$ . Note that while  $V_{out}$  is nearly a constant (the nominal output voltage of the regulator), the input voltage is variable due to the ripple, and thus we can use an average input voltage  $V_{ia}$  instead of the variable voltage, since we are interested in long term heating. It is interesting that as the current goes up, the average input voltage goes down. Thus, the power dissipated in the regulator does not go up exactly in proportion to the current, but at a slower rate.

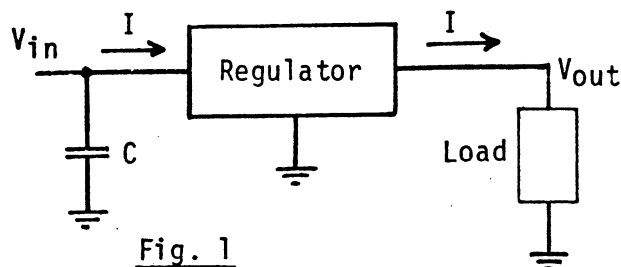


Fig. 1

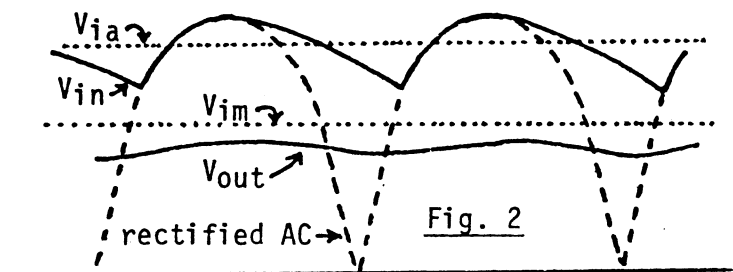


Fig. 2

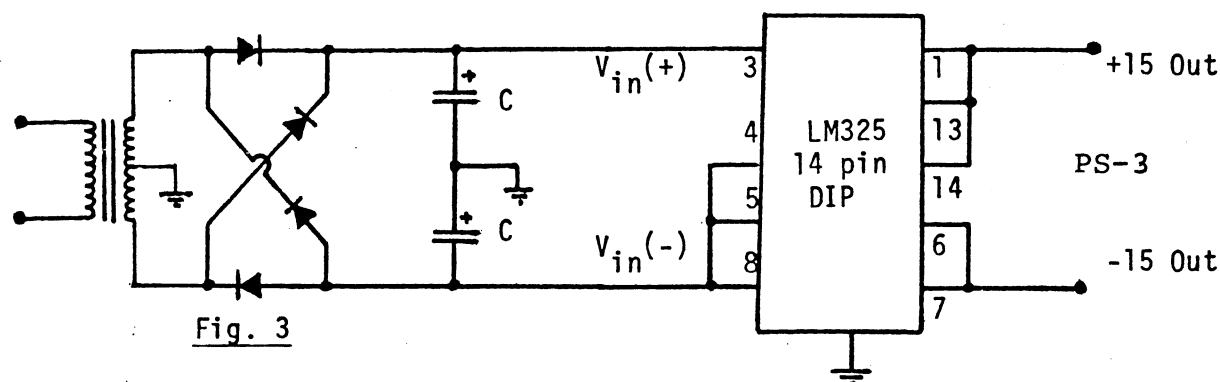
There are really two things which we simply have to achieve with an IC regulator. First, the input voltage must not fall below a certain minimum ( $V_{im}$  in Fig. 2) which is typically about 3 volts above the nominal output, or else the regulator will drop out. Secondly, the IC regulator must not get too hot, or it will shut itself down. As a third, and often relatively unimportant condition, there is a very small ripple on the output (exaggerated in Fig. 2) which is smaller if the input ripple gets smaller. These factors lead to some interesting "trade-offs" in regulator design.

Probably an ideal supply would use a very large filter capacitor and a transformer with a secondary voltage such that the AC rectified peaks went only a few volts above  $V_{im}$ . This would mean that there would only be a few volts across the regulator (low power dissipation and heating) and yet the ripple would be small and well away from the minimum input voltage at all times. However, in practical cases, we often have to use smaller capacitors and transformers with higher output voltages (to allow for changes in actual AC line voltage for example). In this case, the following

considerations become important. If C is made fairly small, the input ripple (and the output ripple - in consequence) increases. However, the "ripple rejection" of IC regulators is excellent, and this is of little concern. The only problem in using a small capacitor is that for larger current demands, the regulator may drop out on the discharge part of the cycle and glitch the supply line, and this is indeed a real problem if it does occur. Here however, we are looking at a small bench supply, and the current demands are low, and held to a maximum value by the dissipation of the regulator, not by the danger of the capacitor becoming discharged to the point where the regulator will drop out. Thus there are two things going for us in a small supply if we use a smaller capacitor. First, the capacitor will be physically smaller and take up less space. Secondly, the average input voltage will be lower and there will be less heating of the regulator.

Let's calculate how small the capacitor can be. We will use the approximate equation  $C = I[\Delta t/\Delta v]$  for a capacitor discharge where I is the current,  $\Delta t$  is the time between charge cycles (1/120 sec. for a full-wave rectifier) and  $\Delta v$  is the full voltage between the peak of the AC cycle and the minimum input voltage  $v_{in}$ . For a small bench supply we might have  $I = 25$  ma max, and the AC peak might be 27 volts while the minimum input voltage for a 15 volt regulator would be 17 volts, making  $\Delta v = 10$  volts. Plugging in these values we get for C the surprisingly low value of 20 mfd., much lower than the 1000 mfd we often put in power supplies.

To test out these values, we constructed two identical supplies based around the LM325 dual tracking  $\pm 15$  regulator. The circuit is shown in Fig. 3 below:



In the unloaded state, the input voltages were about 29 volts for any value of C. By placing 620 ohm resistors on the outputs, the regulator was forced to supply 25 ma of current. We tried values of C of 1000 mfd, 22 mfd, 15 mfd, and 10 mfd. Small drop-outs were seen with the 10 mfd capacitor, but the others were large enough to hold 15 volts at the output at 25 ma. With  $C = 1000$  mfd there was no measurable output ripple. With  $C = 15$  mfd, the output ripple was 0.5 mv, corresponding to a ripple rejection (input to output) of about 74 db. This would not normally present a problem. With  $C = 1000$  mfd, the average input voltage was 25.6 volts, giving a dissipation of 265 mw; with  $C = 15$  mfd, the average input voltage was 22.8 volts, giving a dissipation of 195 mw, a 27% reduction. The "book value" for absolute max dissipation of the LM325 is 650 mw. In terms of actual heat, with  $C = 1000$  mfd, the IC was too hot to touch, while with  $C = 15$  mfd, the IC was hot but you could hold your finger on it indefinitely.

In other bench supplies we have successfully used the three-terminal type regulators in the TO-5 cans such as the 78M15 (+15) and the LM320-15. With the capacitor kept at a minimum value, these will supply from 50 to 75 ma. There is also no reason not to use the larger "power transistor" TO-3 packages, the standard for one amp supplies. These can be soldered in by the two pins, and a separate wire run to a bolt on one of the mounting holes. Without heat sink, you can get up to 200 ma out of these, keeping C to a minimum. [Note that the LM 325 pins given are for 14 pin DIP - many of the application notes give pinout for the TO-5 can. You can also use the type 1568 dual tracking regulator instead of the 325, but this requires some additional passive components.]



1 PHEASANT LANE

June 15, 1979

ITHACA, NY 14850

AN OP-AMP SUPPLY BASED ON  
A 12.6V FILAMENT TRANSFORMER

(607)-273-8030

The need for power supplies (a seeming imposition) has been approached in several ways in these notes. Notes AN-1 and AN-2 discuss basic bench type supplies, which can also be used in medium-to-large sized stand-alone units. A small supply we called the "transferrable power supply" was described in AN-97 and AN-98, and was intended for bench use or small stand-alone units. One problem with these supplies is that they require "op-amp" transformers, a center tapped transformer of about 30 to 40 volts. These can be difficult to obtain, particularly on short notice.

The circuit given in this note has the redeeming feature that it is based around a standard "12.6 volt filament transformer." This type of transformer can be purchased from a "Radio-Shack" type of store just about any time. While these generally are thought to be useless for op-amps (unless you use three of them), we have found that they do work out for light loads.

We found an in/out ratio of about 8.3/1, which gives 12.6 RMS out even if the line voltage drops to 105 RMS. This extra would seem to be an allowance for a large load. Underloaded, the output voltage can be expected to be much higher than 12.6 RMS. At a nominal 117 RMS, the output voltage would be 14.1 RMS. The peak of an RMS voltage of 14.1 occurs at 19.9 volts, and allowing for a 0.7 volt diode drop, you can expect a half-wave rectifier to charge up to 19.2 volts. A +15 volt three terminal regulator such as LM340 or 7815 requires a +17 volt input, so there is not a lot of room here, but with 1000 mfd for the main filter capacitor (See Fig. 1), this should work out.

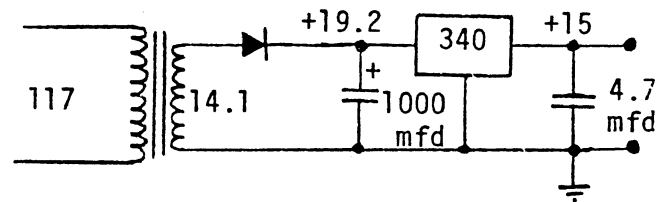


Fig. 1

In fact, we can easily estimate the current that is available here. We can choose a 1 volt ripple (19.2 down to 18.2) at the input of the regulator as the maximum. We then use  $I = C \Delta v / \Delta t$  where  $C = 1000$  mfd,  $\Delta v = 1$  (the ripple) and  $\Delta t = 1/60$  Hz, the time between recharge pulses to  $C$ . Plugging in the numbers gives a current of 60 ma, not bad, and perhaps twice this amount can be expected under favorable conditions.

So much for the +15 supply. We could use the same technique and another transformer to get the -15 supply, but here we want to save one transformer, and will illustrate a useful technique of obtaining a negative voltage from a positive supply. The basic idea has been around for many years, but here we will be using a circuit essentially the same as that of A. Pongsupaht in Wireless World, April 1978, pg. 61. All that is involved is a 555 timer, two diodes, and some R's and C's. The basic idea is to generate an alternating voltage from the +15 supply and then clamp it so that it is always negative, and then rectify it. This is not the best of supplies by any means, but it will work well where supply current requirements and regulation are fairly modest. This is very often the case in simple devices.

In order to understand the full circuit, it is necessary to understand the action of the clamp, and this can be seen in Fig. 2a and Fig. 2b. For the moment, consider only diode D1 and capacitor C1. The input is a square wave that takes on either the value +15 or the value zero. Fig. 2a shows the case when the input is at +15. Clearly a current flows to charge C1, passing through D1 (assumed ideal), and the voltage  $V'$  is zero. Note that C1 and D1 form the classical "diode clamp" circuit. Note that C1 is charged to +15 volts by this process. Now, suddenly the input voltage

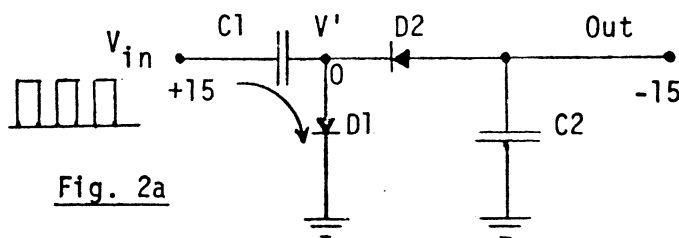


Fig. 2a

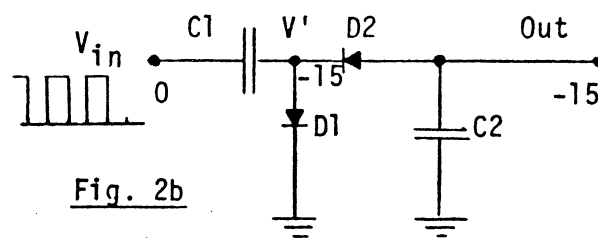


Fig. 2b

changes to zero. The capacitor C1 remains charged immediately after this change, so the result is that V', which was zero, now moves down to -15. When the input again jumps to +15, V' moves to zero again. Here (remember we are ignoring D2 and C2), C1 once charged serves only as a voltage storage, like a little battery, and makes an offset of -15 volts. It is called a clamp because the output cannot rise above zero. Note in particular that C1 does not AC couple the input and D1 does not cut off the positive part (this would leave only -7.5). The reason this does not happen is that for D1 to cut off anything, current must flow, and this charges the capacitor. So much for the classical clamp.

Now, let's pay attention to D2 and C2. Considering the fact that the clamp transforms the waveform to zero and -15, we can view D2 as a simple half-wave rectifier and C2 as a filter. Thus C2 will be charged to -15 and hold it in the absence of a load. Of course, we should observe that C2 is initially at zero, and has to be charged bit by bit to the -15 volt level. Note that any one cycle of the input can supply at most a charge  $Q = 15C/2$ , where  $C = C1 = C2$ .

A full practical circuit, combining Fig. 1 and Fig. 2, and adding the 555 timer is shown in Fig. 3 below:

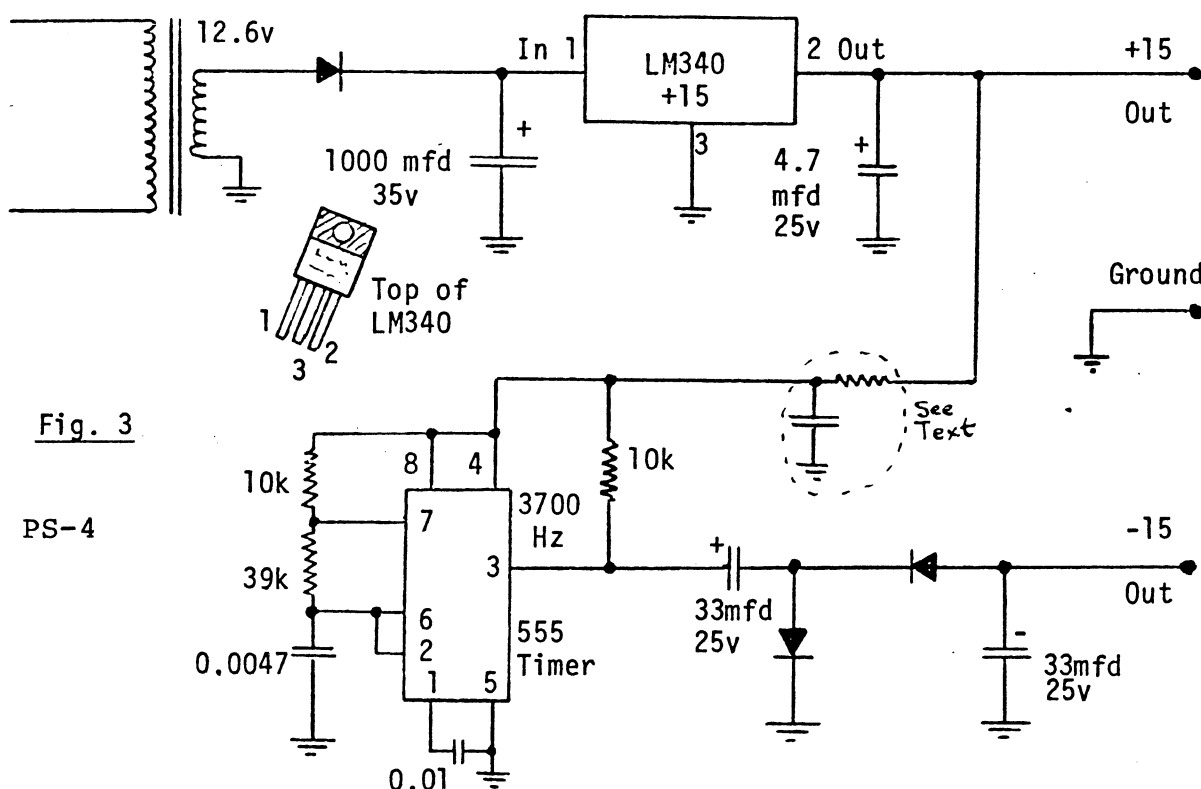


Fig. 3

In some cases, it may be desirable to decouple the +15 line to the 555 timer to prevent the timer from putting noise on the +15 line. Probably 10 ohms in series and 100 mfd from pins 8-4 to ground will do this job. The LM340+15 can be the "T" package as shown, and it will heat sink itself quite well just in free air. You can expect about 4 mv ripple (3700 Hz) for each 3ma of current drawn from the -15 line. This is usually no problem, and 3ma is a good average value for an op-amp. The 33 mfd capacitors can be increased if necessary. The output voltage is not a full -15, more like -14.4 due to diode drop.

What's a crowbar? See EN#35p8 for a simple one; the circuit there can be used on the +15 line by increasing the zener voltage appropriately. A better design for 15V lines uses a 723 regulator, as shown in figure 1. The circuit is shown for protecting +15V, to protect -15V, use the same circuit and exchange +15 for ground and ground for -15V. The circuit is superior to that in EN#35 because should the power supply voltage creep above the trigger level, the change is amplified by the gain of the 723, turning the SCR on harder and faster, decreasing the probability of SCR failure due to slow turn-on. Yes, a surplus 723 will work fine here.

One other typographical error in both ENS-73 and HBSP: The filter capacitor in the 5V supply should be 10,000  $\mu$ F not 1000  $\mu$ F as shown. 1000  $\mu$ F works fine for a light load such as a couple of VCO's and the keyboard interface, but will not supply anywhere near 1.5A as indicated.

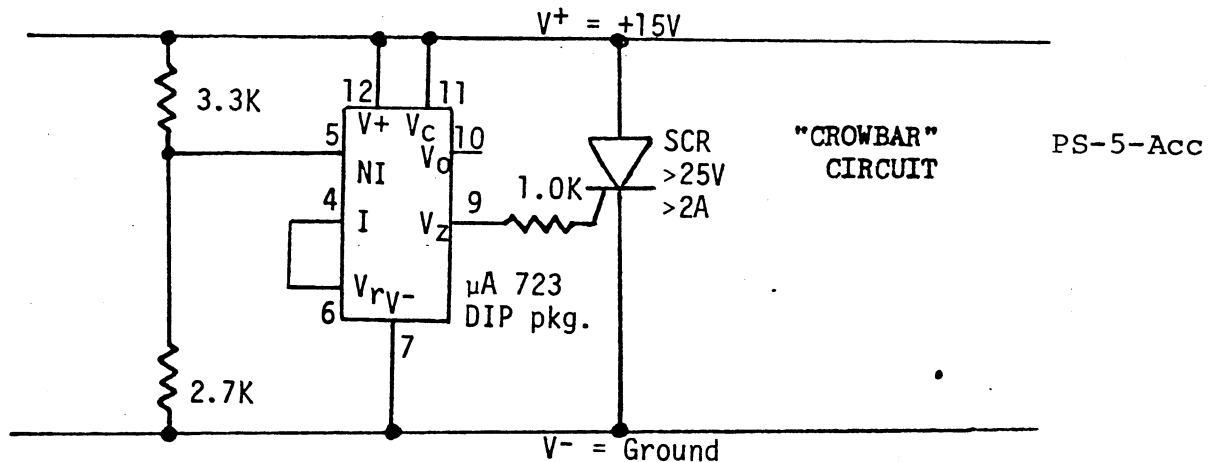
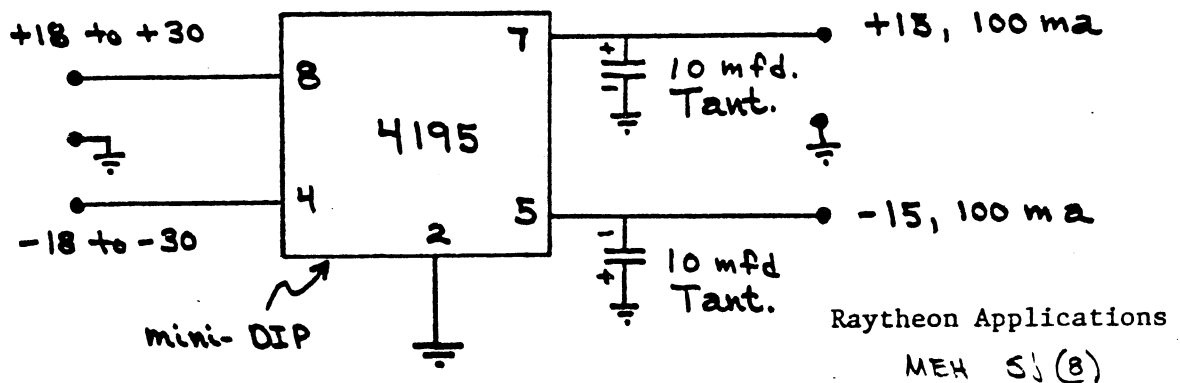


FIGURE 1

# PS-Extra Circuit





## CONTROLLER INTERFACES

Controller Interfaces: Here we will present two controller interface circuits. The first is a digital keyboard interface, and will be presented by reprinting in entirety the original presentation from EN#68. The second version is an analog interface, which also works with a keyboard, but will also work with some other controllers such sample-and-holds and sequencers, and even ribbon controllers played in a discrete-tap mode. Both units are roughly equivalent in cost and construction time. The digital unit is preferred if only a keyboard is to be used, while the analog unit makes a good second interface. However, the keyboard wiring is different so you can't easily switch the keyboard from one unit to the other. The controller interface options are:

| <u>OPTION</u> | <u>SOURCE</u> | <u>DESIGNER</u> | <u>APPROX. COST</u>            | <u>NOTES</u> |
|---------------|---------------|-----------------|--------------------------------|--------------|
| CI-1          | EN#68         | Hutchins        | \$20*                          | Digital      |
| CI-2          | EN#45         | Hutchins        | \$20*                          | Analog       |
| CI-3-Acc      | MEH 7b        | Hutchins        |                                | Ranging and  |
| CI-4          | EN#124        | Gualtieri       | \$100 with<br>surplus keyboard | Scaling Unit |

THE ENS-76 HOME-BUILT SYNTHESIZER SYSTEM - PART 4: -by Bernie Hutchins

In this installment we will look at our choice for a digital keyboard. The circuit we have chosen is essentially "Design Example Number 2" on pages 4, 6, and 7 of EN#62. This circuit serves to wire up the keyboard and interface it with the other modules. It provides a control voltage representing the key that is down, and also gate and trigger signals for controlling envelope generators. The circuit gives one note at a time, and always gives the lowest note pressed if more than one is down. We have found this circuit to work very well, and a number of readers expressed a preference for the "Position Rule" (lowest key down) even though the circuitry is a little more complicated. This position rule is exactly what you get from the standard voltage divider keyboard and apparently many players rely on it for their playing style. For example, they can trill rapidly by holding one key down (the upper one of the trill), and rapidly tapping the lower one. The circuit given here performs identically with the earlier ENS-74 analog interface. It would be difficult to tell which one you were using except the digital circuit has the advantages of digital accuracy and stability. We had a number of "debounce" schemes in mind for this circuit. However, the latching of the circuit seems to provide most of the debouncing required. One additional debounce technique will be given as an option. Finally, in working with this we had a chance to look again at the mechanical bouncing that proved a problem in the ENS-74 system. It was also present here, but we were able to get rid of it by mechanical means. Persons still using the ENS-74 analog interface will want to look at the comments by Ian Fritz last issue, and also the comments in this presentation.

The circuitry is presented in four parts (Clock, Scan-Counter, Scan-Control, Latches; Scan-Multiplexer; Gate and Trigger Circuitry; and D/A Converter). These circuits appear in Fig. 1, Fig. 2, Fig. 3, and Fig. 4 respectively. It may aid your understanding if you make photocopies of these figures and paste them on the wall in their proper positions. After presenting the circuit we will talk about wiring and adjusting the keyboard, and then discuss a few options.

## CLOCK, SCAN COUNTER, SCAN CONTROL, AND LATCHES

The circuitry to be discussed in this section is shown in Fig. 1. IC-1 is a 555 timer used as the basic clock and operates at about 15 KHz. The clock drives IC-2 and IC-3 which together form a 6-bit binary counter. Note that the clock will also be fed to a "Trigger blanking counter" as noted on the diagram. The 6 bits are denoted A, B, C, D, E, and F. When latched, A will be the least significant bit (LSB) while F will be the most significant bit (MSB) in the digital word. The function of IC-3 could be performed with a 7473 or another 7493 if this is more convenient. IC-6 is an 8-input NAND gate (used here as a 6-input since two inputs are held high at +5). Note that the counter runs continually overflowing to zero about every 4 milliseconds. The counter thus hits binary 63 (111111) once every 4 ms and the output of IC-6 goes low resetting the RS flip-flop (IC-7a, b) so that the output of IC-7a is high. If a select pulse (high) arrives from the multiplexer, this will pass through IC-7c, and appear as a low at the output of IC-7c. This low then cycles to the input of IC-7b, flipping the RS flip-flop back so that the output of IC-7a is low. This in turn blocks what's left of the select pulse that is trying to get through IC-7c. The output of IC-7c is thus just a very short negative going pulse, and this is inverted by IC-7d and causes the latches (IC-4 and IC-5) to latch the digital word corresponding to the select pulse from the multiplexer. We used 7475's for the latches because they were available. Probably a single 74174 six-bit latch would work just fine as well. We show in Fig. 1 the D/A and the multiplexer as dotted boxes. With these functioning, we have a digital keyboard with voltage hold, but without gate and trigger. Note that since the RS flip-flop resets at count 63 (above the top of the 61 note keyboard), and then flips back as soon as a note is found, that only the lowest note is latched. This is the way that the position rule is implemented. Also, when no key is pressed down, there is no select pulse emitted from the multiplexer. This means that the last note is held after all keys are released.

FIG. 1  
CLOCK, SCAN COUNT  
IC-1  
IC-2  
IC-3  
IC-4  
IC-5  
IC-7a,b,c,d  
IC-7430

FIG. 2  
MULTIPLEXER

Note: Add about three 0.01 capacitors to this circuit to bypass the power supply lines.

**Note:** Add about three 0.01 capacitors to this circuit to bypass the power supply lines.

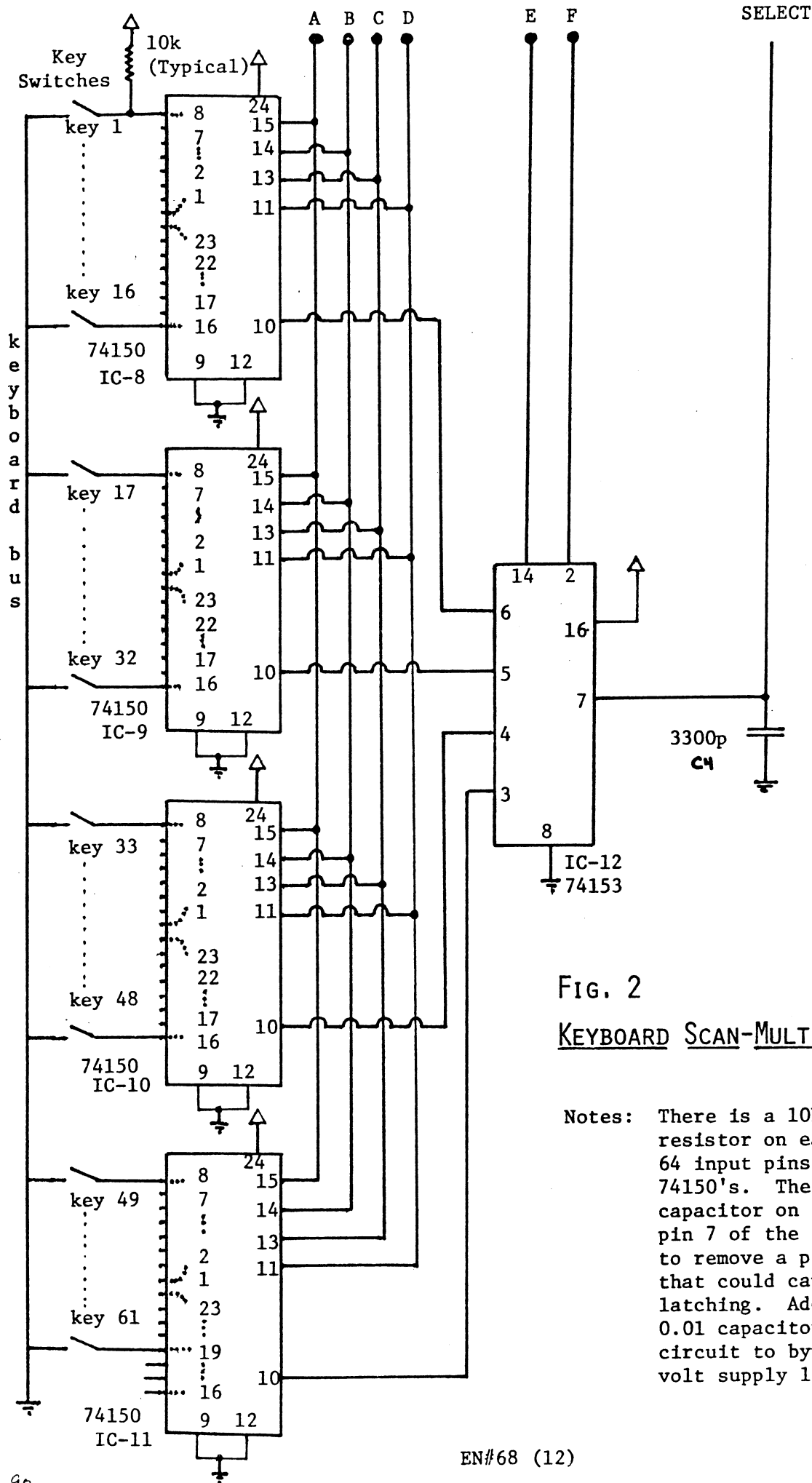


FIG. 2

# KEYBOARD SCAN-MULTIPLEXER

Notes: There is a 10k "Pull-Up" resistor on each of the 64 input pins of the 74150's. The 3300p capacitor on the output pin 7 of the 74153 is used to remove a possible glitch that could cause false latching. Add about five 0.01 capacitors to this circuit to bypass the +5 volt supply lines.



## KEYBOARD SCAN-MULTIPLEXER

The keyboard scan-multiplexer circuit is shown in Fig. 2. Consider first the two address lines E and F which enter IC-12. These are the two most significant bits from the scan counter. They count 0, 1, 2, and 3. When the E-F lines are 0-0, the line from pin 10 of IC-8 is accepted into pin 6 of IC-12. The E-F lines will not change until the A-B-C-D lines have counted from 0-0-0-0 to 1-1-1-1. Thus, the A-B-C-D lines cause IC-8 to scan keys 1 to 16 in series. When the A-B-C-D lines change from 1-1-1-1 to 0-0-0-0, the E-F lines change from 0-0 to 1-0. This causes IC-12 to accept data from pin 10 of IC-9 rather than pin 10 of IC-8. The A-B-C-D lines again count upward from 0-0-0-0 to 1-1-1-1, and are now scanning keys 17 to 32. The process continues in the same manner.

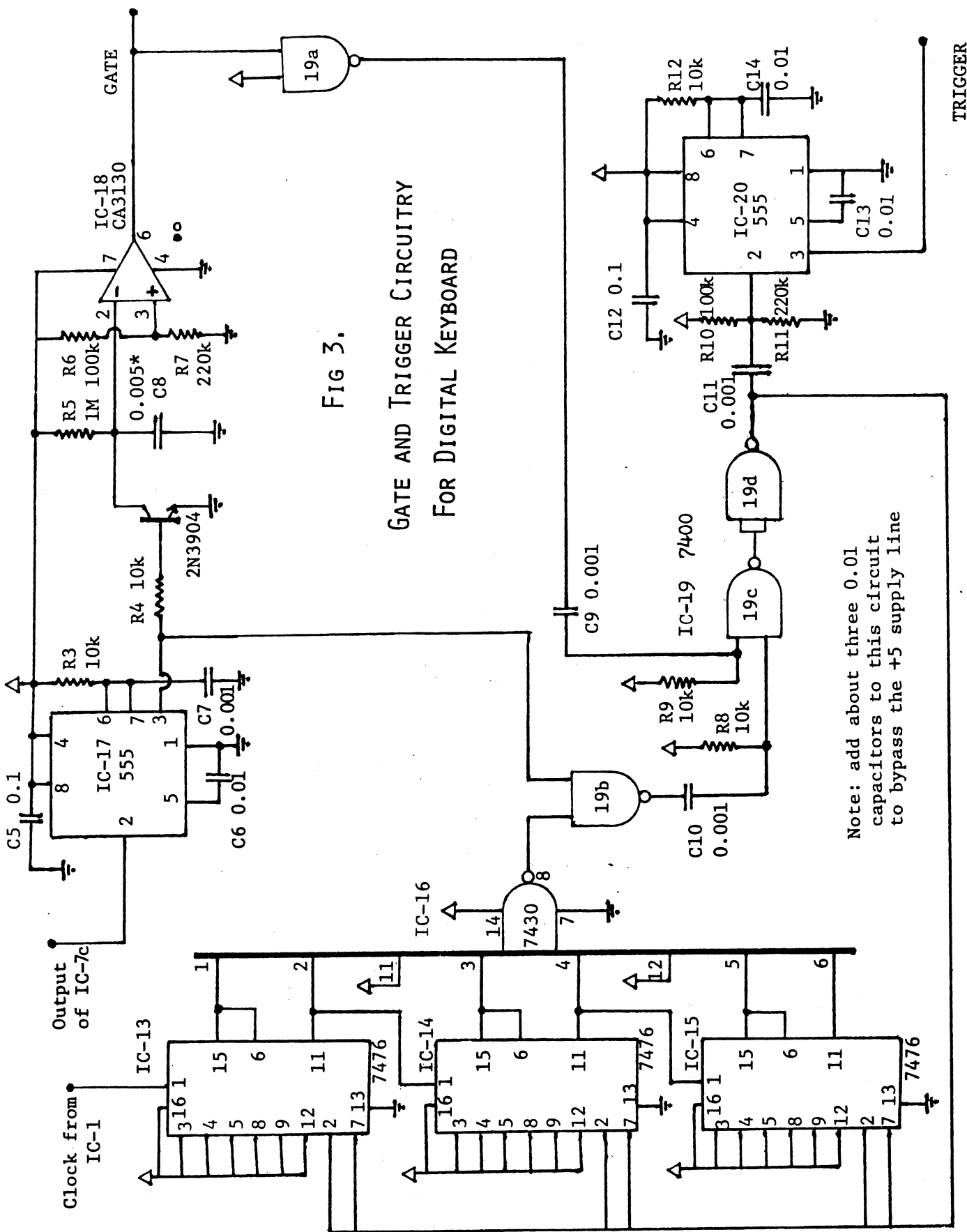
Thus it can be seen that IC-8, IC-9, IC-10, and IC-11 form 16-to-1 data selectors. These are cascaded into IC-12 which is a 4-to-1 data selector. The combined setup is then a 64-to-1 data selector which scans the keys of the keyboard looking for one that is down. If we had a 64-to-1 data selector, we would use it, but what we have to settle for is some sort of cascade arrangement. Some users may want to use eight 74151's cascaded into a ninth 74151 to form the 64-to-1. The 74151 is an 8-to-1 data selector. We are just looking for a scheme that looks for a low key, and gives a high output at the select terminal whenever a key is down.

Note that all 64 of the inputs to the four 74150's are pulled up with a 10k resistor as per the example shown on pin 8 of IC-8. These resistors can actually be any convenient value in the range of 2k to 10k. Don't allow pins 16, 17, and 18 of IC-11 to float. Either use pull-up resistors, or hard wire them to +5. Pin 18 may be used for a 62nd note, but don't try to use pins 17 and 16, as this confuses the scan logic.

## GATE AND TRIGGER CIRCUITRY

The gate and trigger circuitry is shown in Fig. 3 on page 14. The output of IC-7c is the only signal that drives this section, except for the basic clock from IC-1. The output of IC-7c is a short negative going pulse that emerges about every 4 ms whenever a key is down. We saw in Fig. 1 that the timing of this pulse was such that the proper digital word was latched. Here we are first concerned with the fact that they emerge every 4 ms. IC-17 standardizes the negative going transition to a positive pulse of 10 microseconds duration. This pulse discharges C8 through the 2N3904 transistor. C8 then starts to recharge toward +5 volts through R5. If another pulse arrives within about 4 ms, the capacitor discharges, and never reaches the reference voltage of 3.44 volts set by R6 and R7. The gate is thus high as long as pulses emerge every 4 ms. If all keys are lifted, no pulses emerge from IC-17. The capacitor then charges above 3.44 volts and the output of IC-18 goes low. The gate is thus seen to be formed by IC-17 and IC-18 which are used as a retriggerable monostable. Probably a retriggerable monostable like the 74122 could be used. We used the IC-17 and IC-18 version because we needed the standardized pulse, and because the CA3130 offers a well defined (zero or +5) gate condition. Note that we show C8 as 0.005 mfd. Our circuit actually has a 0.004 mfd capacitor. This sets the monostable for pretty close to 4 ms, and we felt that circuit tolerances might give glitches in the gate for some components. Thus we suggest you start with 0.005 as shown and unless there is some reason to shorten it (as there might be if the digital word is being transferred under gate control), leave it at 0.005.

Next we follow the rising edge of the gate through inverter IC-19a and C9. The negative going edge out of IC-19a causes a short positive pulse at the output of IC-19c and a negative pulse at the output of IC-19d. This pulse out of 19d does two things. First it triggers IC-20 to give a +5 volt trigger for external use. This pulse has a



0.1 ms duration. The second thing that happens is that the trigger blank counter (IC-13, IC-14, and IC-15) is reset to 111111. This counter then advances at the same rate as the scan counter since they have a common clock. Thus we see that the rising edge of the gate causes an external trigger to emerge. If the key that was pressed down remains down for more than 4 ms, a second pulse will emerge from IC-17. However, since the trigger blank counter has advanced exactly 64 counts after it was reset, it is once again in the 111111 state, IC-16 is satisfied (low) and this pulse from IC-17 is blocked from passing through IC-19b. This will happen as long as the same key remains down. If the key is released, the gate goes away, and that's all. If the user then presses another key, the leading edge of the gate provides the trigger and synchronizes the trigger blanking counter exactly as in the original case. If the second key is the same as the first, the circuit synchronizes, but would have been in synchronization anyway. The final case to consider shows why the trigger blank counter is necessary, since up to this point, we could have just used the leading edge of the gate as a trigger. Suppose you hold one key down. The trigger blank counter synchronizes. If this key is held down and a second one is pushed down at the same time, a scan pulse will emerge earlier than 64 counts after synchronization if the second key is below the first. [If it is above the first, it will be ignored.] Since the trigger blank counter was set in synchronization with the first key, the output of IC-16 will not be low when the new pulse emerges, this will pass through IC-19b, C10, IC-19c, and IC-19d, and cause a new trigger. It will also synchronize the trigger blank counter to the new lower key position.

## DIGITAL TO ANALOG CONVERTER

The basic theory behind the D/A converter in Fig. 4 (page 16) was given in the "Back To Basics" section in EN#67. It is simply a R-2R ladder D/A converter. There is little more to say here. Note that it was necessary to use CMOS to drive the D/A R-2R ladder rather than TTL since the CMOS levels are better defined. There are many ways to make this conversion. We chose to simply use a 7404 hex inverter followed by a 74C04 hex inverter (IC-23 and IC-22 respectively). Both these chips are powered by +5 volts. The R-2R ladder we used was composed of 332k 1% resistors with 332k being 2R and the 166k resistors were formed with two 332k in parallel. The exact value of these resistors is not too critical. They should be in the range of 50k to 1 meg, and matched to 1% or better. We choose the CA3130 op-amp due to its high impedance input. This allows us to choose a wide range of possible values for R and also choose values that are quite high so that the impedance of the CMOS inverters (several hundred ohms) is negligible. Note that the CA3130 is run on +15 volts since +5 is not enough for a five octave keyboard. Note that the op-amp provides a small gain (nominally 1.06666..) for a +5 volt supply. Depending on the exact voltage of your +5 volt supply, you may have to adjust the value of R14 up or down. TP-1 can be easily adjusted once the keyboard is finished since all you have to do is press keys an octave apart and see if the voltage changes by one volt or not.

We found that the 1% resistors used worked quite well. For best results however you may want to use 0.1% resistors. Since the exact value is not very critical, you can just watch for some usable value with 0.1% tolerance to appear at surplus prices. If you have a DMM, you can select resistors, or see the matching technique given in EN#63 by Carl Fravel. If you have a bunch of 100k 1% resistors for example, you may be able to select them to a 0.1% match. Or you can decide to make them all into 101k resistors. To do this, you would just measure the resistance and add on approximately the right 5% resistor to make 101k. For example, if it measured 100.2k, you could add a series 820 ohm resistor. This procedure will get you plenty of accuracy for a keyboard application (equivalent to nearly 0.1% accuracy in the resistors).

Capacitor C16 compensates the CA3130, and C15 adds some additional stability to the D/A voltage. If you have vary large values of R, you may need to decrease C15 to avoid unwanted portamento.



to attach the keyboard wiring, but for a permanent setup, it may be easiest to just solder the ribbon cable to the data selector chips. In our setup, the data selector chips are mounted on the foil side of the board, pull-up resistors are mounted under the chips (the 74150 is a big 24 pin chip), and the wires of the ribbon cable are soldered to the "shoulder" of the IC pins. The diagram in Fig. 5 should give you the idea. This makes it possible to keep things compact, and the selector chips serve as the receiving point from the actual keyboard. It is possible to do this with 1/4 watt resistors, but it will be very tight. It would be better to use 1/10 watt

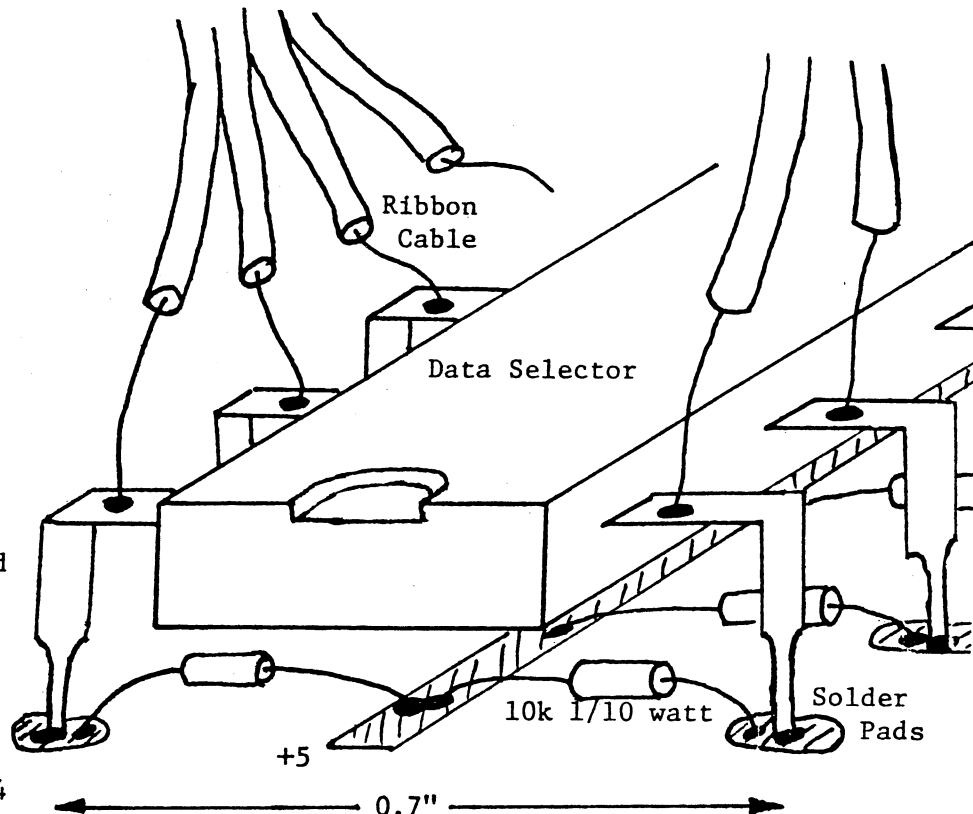


FIG. 5 WIRING OF DATA SELECTORS

resistors for this. Another thing that we considered but did not try was to use a chunk of velofoam (the black foam material that CMOS is shipped in) under the data selector chips and in contact with the selector pins and with a +5 volt strip near the center of the chunk. A couple of tests indicated that the resistance might be about right for this. If so, it could replace all the pull-up resistors and probably would be easier to install. It is also possible that voltage gradients in the material might be such that when one pin is pulled down, other nearby might be also, so some testing would be indicated.

## MECHANICAL ADJUSTMENT OF KEYBOARD

Most of the keyboards you will be likely to obtain will be the standard Pratt-Read type which are basically electronic organ keyboards with a single bus and one switch per key. The switching mechanism consists of spring wires which are forced against a bus by an activation post connected to the key. This mechanism is shown in Fig. 6 (looking from under the keyboard). These keyboards generally work fine, but there is a problem with them which involves mechanical bouncing of the spring wires. This is not the same thing as the bouncing of the switch contacts that is familiar. Here, during release the wire is allowed to spring back from the bus and is then stopped by the support structure of the key. It then springs back and strikes the bus again causing triggering problems. In the present design, the spring wires are electrically common with the pins of the data selector which are held

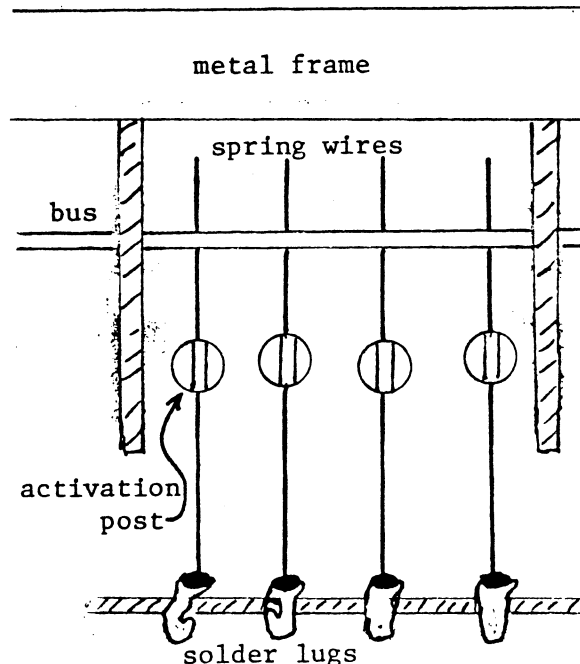
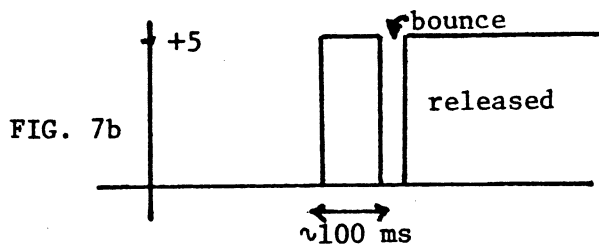
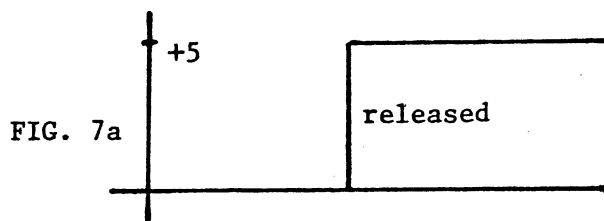


FIG. 6 KEY SWITCH WIRES

at +5 volts by the pull-up resistors. When these wires contact the bus which is grounded, they go to ground potential. Thus, the release of a key should result in a key voltage as shown in Fig. 7a. In many cases, this is just what you do get. With some keyboards however, there may be bouncing problems when the keys are released very rapidly. For example, when the key is held down and the finger is slid off the end, the key springs back as rapidly as possible. If bouncing is to be a problem, this will show it. In the bouncing case, the key voltage is as shown in Fig. 7b. The bounce causes a retrigger.



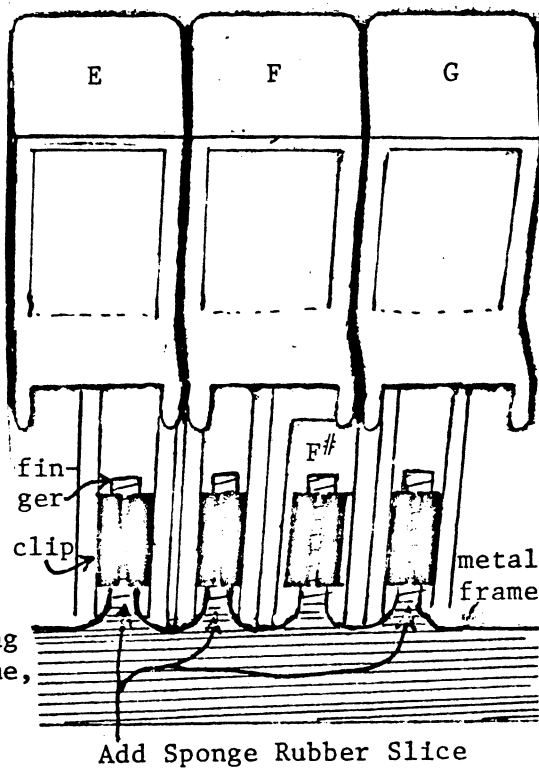
Fortunately, there is a fairly simple solution that will most likely work to prevent this bouncing. Three things can be done mechanically. First, the spring wires can be bent up away from the bus slightly so that when they snap, they don't reach the bus. They can be bent with a slight pressure of the finger applied just above the activation post. I suggest testing all the keys by snapping them and marking those that bounce with a crayon. In general, there is no problem with the black keys (you can't really snap them anyway!). Bend up the wires of those that cause problems and test again repeating the bending as is necessary. This should cure the problem. You will have to push the keys down a little further than before however. For this reason, you may want to try two other things first even though it does not seem that either will completely cure the problem.

The second thing to try (that is, the first of the partial cures) is to add some sponge rubber so that the keys do not strike their supports so hard when they snap back. Fig. 8 shows the underside of the keys from the front. The solid black areas are metal clips that fasten to the core of the key while the cross hatched area is the metal support frame which has fingers bent at 90° which restrain the motion of the clips (and thus catch the key when it returns. These fingers are covered with a rubbery material, but a small slice of foam rubber inserted between the finger and the clip will soften things up a lot. It will also decrease the distance between the spring wire and the bus (since the key is lowered), and this may be the reason why the sponge rubber by itself does not cure the problem.

The final thing that you can consider is apparent if you again look at Fig. 6. The spring wires are about 1/4 inch longer than they need to be. Since the problem is one of inertia of the moving wire, you don't have to be a mechanical engineer to see that the wires aren't going to bounce so much if you shorten them. If you're like me, you hesitate a moment before taking a wire cutters to something you paid quite a bit of money for. I did cut them and didn't find that it did any harm. I also don't believe it solved the problem although it is hard to tell because in the process of cutting the wires you bend them a little one way or another.

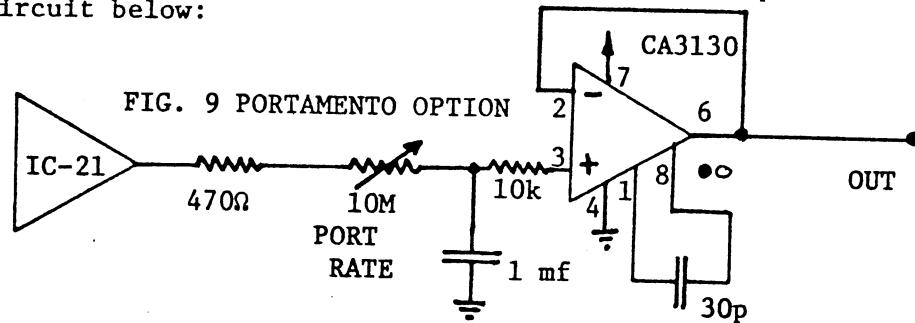
The above steps should serve to cure any bounce problems you may have. We also feel that what we found with this system also applies to the ENS-74 keyboard circuit as well.

FIG. 8 UNDERSIDE OF KEYBOARD

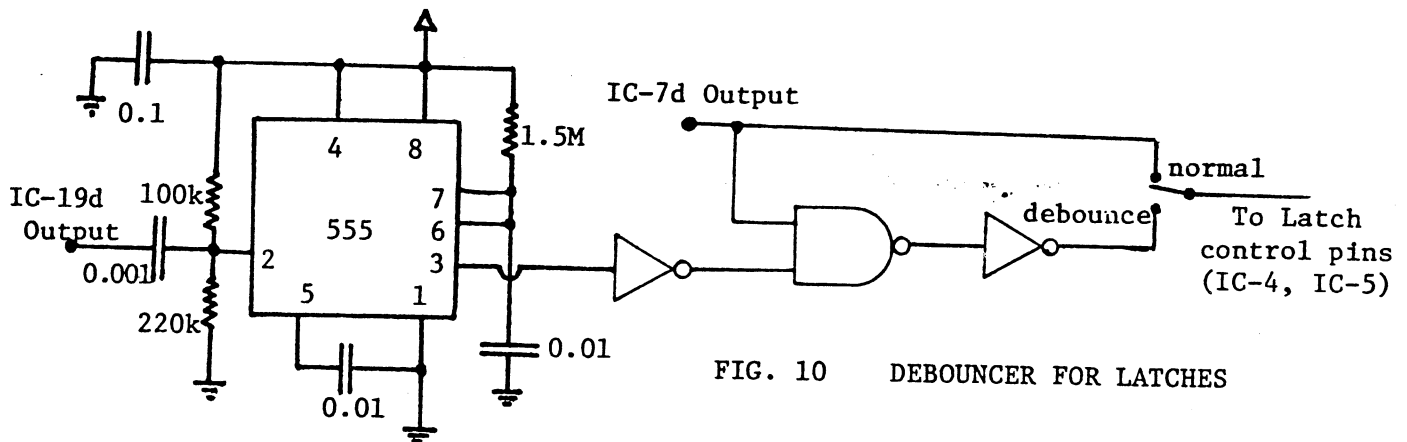


## OPTIONS TO ADD TO THE KEYBOARD CIRCUIT

You will most likely want to add a portamento circuit to your keyboard.. This is a very easy thing to do as long as you just want standard exponential portamento. We suggest the circuit below:



The second option to be discussed is a "debounce" option. Any mechanical switch contacts make and break several hundred times before finally settling. The bouncing may go on for several milliseconds. This has nothing to do with the problem with the bouncing spring contacts - it is present in any mechanical switch you have. The latching effect of the present keyboard circuit does much to prevent debounce problems since the circuit responds to the first "make" of the switches. The problem comes up when more than one key is down. What should happen when two keys are down? Logically you should get two notes (it is a keyboard after all), but we are just using it with a one note instrument. Thus, we employ some sort of rule (here the position rule) that tells us what is to happen - the lower of the two notes sounds. Thus, we know what will happen if one key goes down and another follows. The remaining problem is to consider what will happen if two keys go down at the same time so that both are bouncing at the same time. The latching circuit will jump from one to another. Well, bearing in mind that the two note problem on a one note instrument is nonsense at best, we would like something to happen that does not call our attention to this unfortunate situation. What we can do is use a circuit that blocks relatching for a certain "debounce" period after a latch command arrives. Such a circuit is shown below:



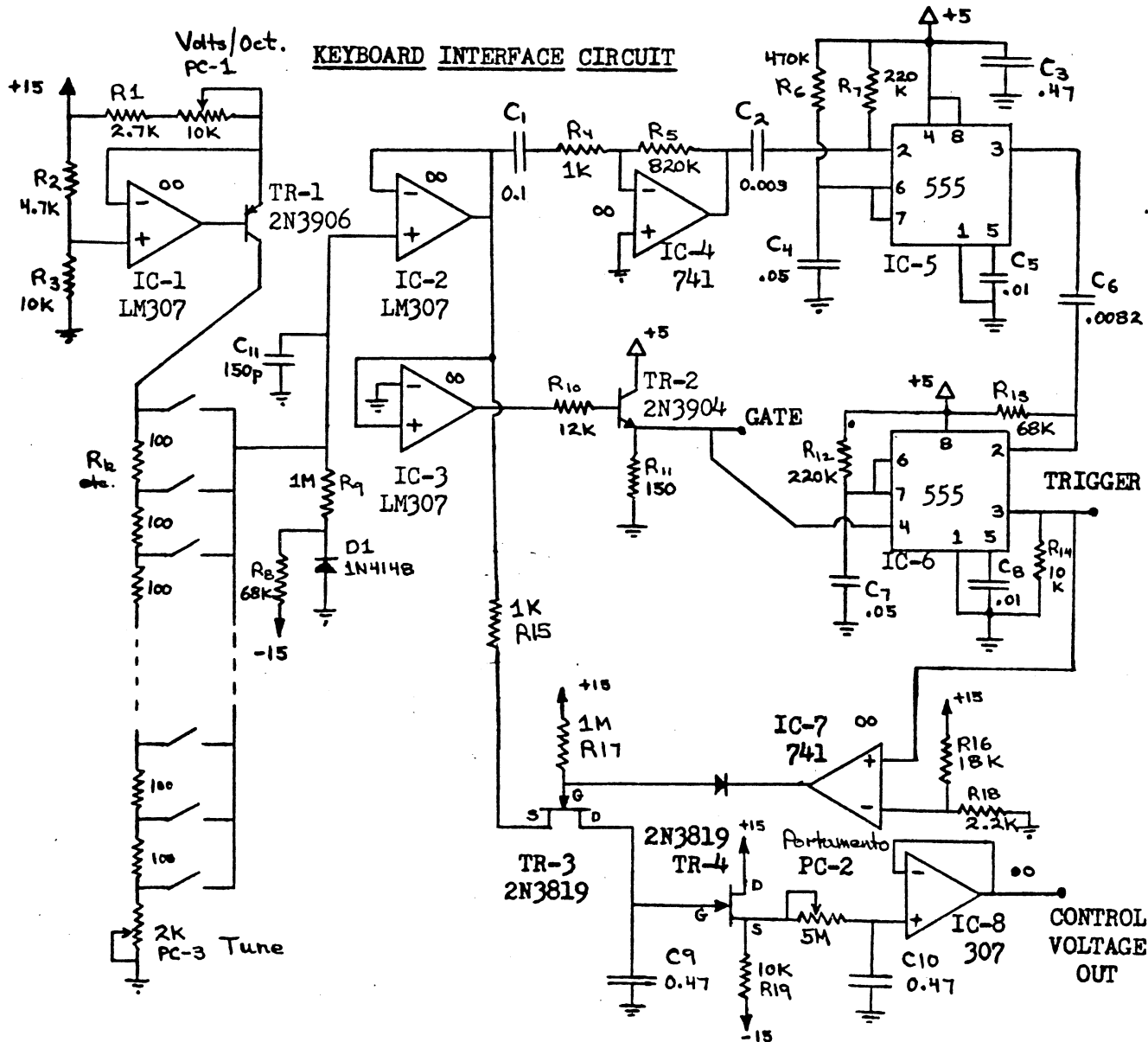
It is difficult to say whether this circuit or some variation need be included. Also, some work could be done to see what the best time constant for the monostable is. This can be changed by changing the 1.5M resistor up or down. As shown, the monostable goes high for about 15 ms and prevents relatching during that time.

Another "debounce" idea that you may want to consider is to simply increase the value of C15 up toward the point where portamento becomes noticeable, and then cut it about in half. There are many debounce schemes that you may try. It is probably best to make them switched options (as in Fig. 10) until you see if they really help.

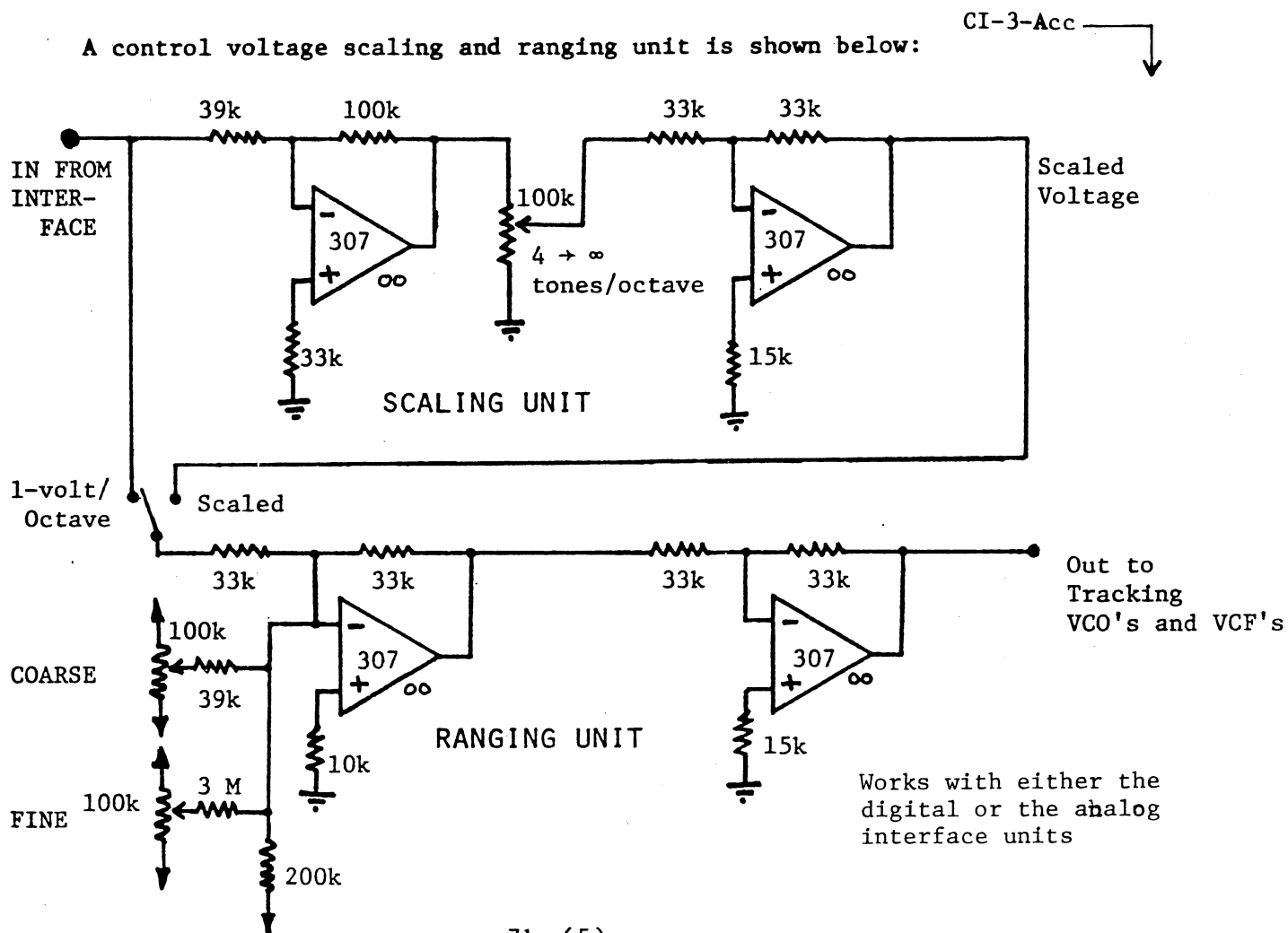
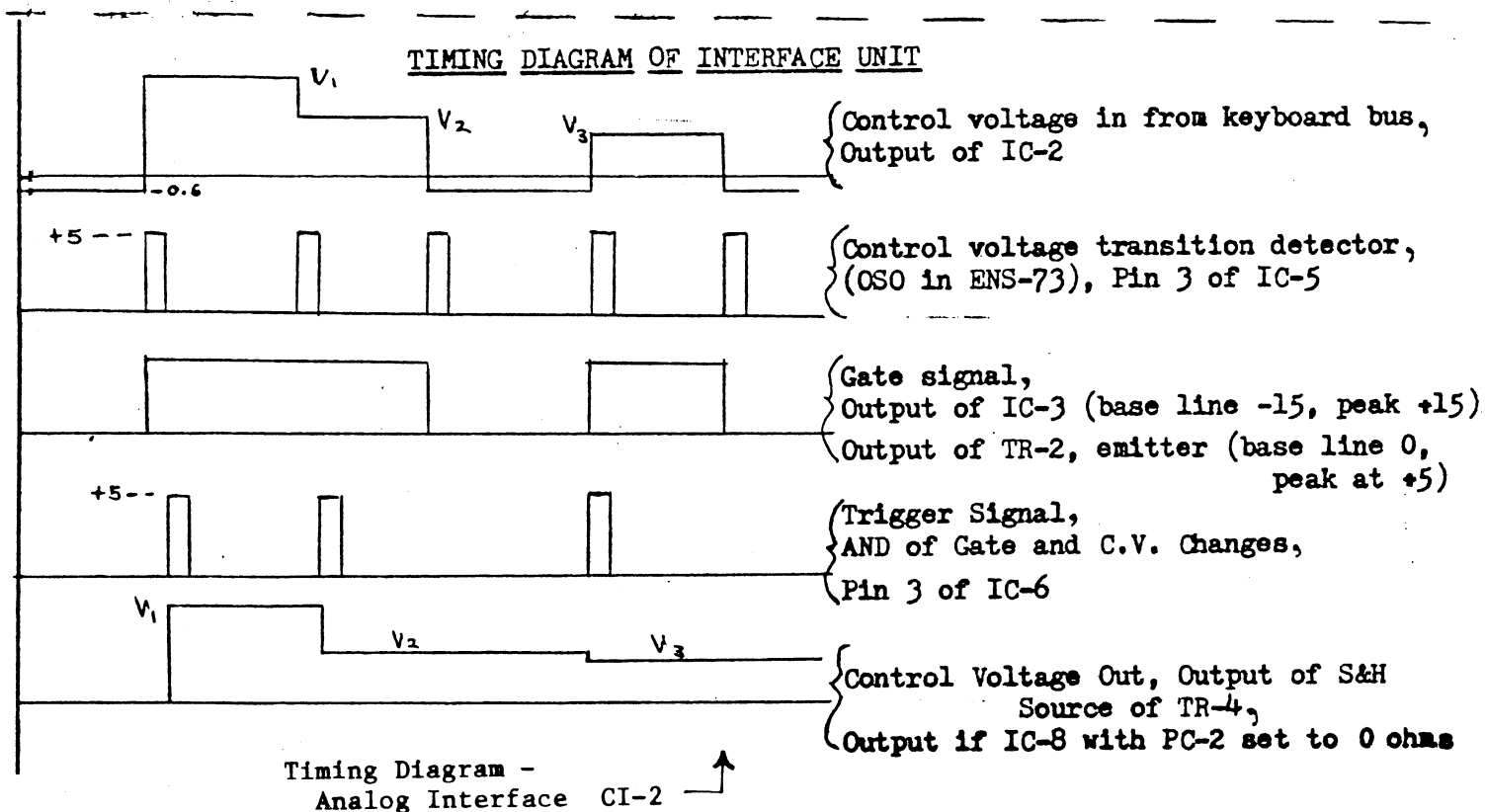
## Controller Interface CI-2 (Analog)

The controller interface in the diagram below is reprinted from EN#45. IC-1 and TR-1 form a current source driving the string of 100 ohm resistors on the keyboard ( $R_k$ ). The resistors are shown as 100 ohms, but they could be any value between 10 ohms and 2k ohms if the current source is adjusted so that each 12 resistors develop a 1 volt drop. IC-2 buffers the keyboard voltage while IC-3 and TR-2 derive a gate (a key is down). IC-4 is a differentiator (to detect a changing voltage) while IC-5 and IC-6 sequence triggering. IC-7 triggers the sample-and-hold switch TR-3, charging C9 when triggered, and the voltage on C9 is buffered by source follower TR-4. A portamento function is derived using PC-2/C10/IC-8. In a modern design, you could replace IC-2 and IC-8 with a BiFET amplifier (e.g., LF351), and the source follower TR-4 could also be replaced with an LF351 as a follower.

The operation of the unit can be understood from the timing diagram on the next page. Pin 3 of IC-5 goes high on every transition of control voltage (every change or key down, even if the first key is not lifted), but only when the gate is down will an actual trigger (or retrigger) be generated to resample the voltage and to trigger an external envelope generator. This prevents sampling of no-key down transitions.







This controller interface involves a complete surplus (ARP) keyboard which has attached a digital interface unit. This unit will be included in all 1982 issues of the EBG&PCC, but will be dropped for 1983 unless it is established that the surplus stock is still available.

*These are probably gone, but we  
will continue this for those  
who can still find one*

*April 1984*

# THE ARP™ SURPLUS THREE-OCTAVE KEYBOARD.

-by Dev Gualtieri

I recently purchased a surplus three-octave ARP keyboard, which was advertised on page 141 of the December 1981 issue of Radio-Electronics from Computers, Peripherals, Unlimited (formerly, CFR Associates), Box 204, Newton, NH 03858 (Phone 617-372-8637). I received it just a week-and-a-half after mailing out my order, and I was surprised to find that it contained an integral decoder board. My surprise turned to pleasure when I saw that all the ICs on this board (Board E, ASSY 7203101, FAB 5402401) were clearly marked TTL packages, and that the board terminated in a sixteen-pin DIP socket and a fourteen-pin DIP plug. I immediately attached a five-volt power supply to the power pins of one of the ICs and went at the DIP plug and the DIP socket with my TTL probe.

What I found was a fully functional monophonic keyboard decoder, operating from five-volts at (estimating from the package count) about 200 mA, which outputs and holds a digital word corresponding to the lowest of the keys pressed and generates gate and trigger signals. Table I and Table II show the pin-outs of the DIP plug and socket, and Table III shows the digital codes generated.

As can be seen from the Tables, a simple shunt header placed in the DIP socket will transfer the key-pressed digital signals to the DIP plug, so that the plug will have all the relevant signals. The DIP shunt will connect the socket pins 1-16, 2-15, 3-14, 4-13, 5-12, 6-11, 7-10, and 8-9, so that the pin-out on the DIP plug will include 9 - "A", 10 - "B", 11 - "C", 12 - "D", 13 - "E", and 14 - "F".

The digital coding is not monotonic, and simply connecting a D/A converter to lines "A"- "F" will give an unusual (since we're talking about music, it's not really wrong) scaling. "A" and "B" encode the octave, and "C"- "F" encode the note in the octave. Since "C"- "F" can potentially encode sixteen states, and there are only twelve notes, there is a four state gap between the B and C keys.

If the keyboard is to be connected to a microprocessor, as mine will be, software will rectify this problem easily. This particular coding scheme will even make some of the programming easy. If you want an analog output, two D/A converters are required, one for the octave lines "A"- "B", and one for the note lines "C"- "F". The

analog outputs of these D/A converters would then be summed to give the correct monotonic output scaling across the keyboard. This seems at first to be a nuisance, but it actually turns out to be an advantage. Two inexpensive D/A converters, such as those with eight-bit resolution, can be used, with careful trimming, to give the same accuracy of tuning as a more expensive converter, such as one with ten bit resolution.

Table I. Pin-out of DIP plug.

| <u>Pin</u> | <u>Function</u> |
|------------|-----------------|
| 1          | GROUND          |
| 2          | GROUND          |
| 3          | +5 VOLTS        |
| 4          | +5 VOLTS        |
| 5          | SOCKET-2        |
| 6          | GATE            |
| 7          | TRIGGER         |
| 8          | SOCKET-15       |
| 9          | SOCKET-14       |
| 10         | SOCKET-13       |
| 11         | SOCKET-12       |
| 12         | SOCKET-11       |
| 13         | SOCKET-10       |
| 14         | SOCKET-9        |

Table III. Digital output codes.

| <u>Key</u> | "A" | "B" | "C" | "D" | "E" | "F" |
|------------|-----|-----|-----|-----|-----|-----|
| CO         | 0   | 0   | 0   | 0   | 0   | 0   |
| C#0        | 0   | 0   | 0   | 0   | 0   | 1   |
| DO         | 0   | 0   | 0   | 0   | 1   | 0   |
| D#0        | 0   | 0   | 0   | 0   | 1   | 1   |
| EO         | 0   | 0   | 0   | 1   | 0   | 0   |
| FO         | 0   | 0   | 0   | 1   | 0   | 1   |
| F#0        | 0   | 0   | 0   | 1   | 1   | 0   |
| GO         | 0   | 0   | 0   | 1   | 1   | 1   |
| G#0        | 0   | 0   | 1   | 0   | 0   | 0   |
| AO         | 0   | 0   | 1   | 0   | 0   | 1   |
| A#0        | 0   | 0   | 1   | 0   | 1   | 0   |
| BO         | 0   | 0   | 1   | 0   | 1   | 1   |
| C1         | 0   | 1   | 0   | 0   | 0   | 0   |
| C#1        | 0   | 1   | 0   | 0   | 0   | 1   |
| D1         | 0   | 1   | 0   | 0   | 1   | 0   |
| D#1        | 0   | 1   | 0   | 0   | 1   | 1   |
| E1         | 0   | 1   | 0   | 1   | 0   | 0   |
| F1         | 0   | 1   | 0   | 1   | 0   | 1   |
| F#1        | 0   | 1   | 0   | 1   | 1   | 0   |
| G1         | 0   | 1   | 0   | 1   | 1   | 1   |
| G#1        | 0   | 1   | 1   | 0   | 0   | 0   |
| A1         | 0   | 1   | 1   | 0   | 0   | 1   |
| A#1        | 0   | 1   | 1   | 0   | 1   | 0   |
| B1         | 0   | 1   | 1   | 0   | 1   | 1   |
| C2         | 1   | 0   | 0   | 0   | 0   | 0   |
| C#2        | 1   | 0   | 0   | 0   | 0   | 1   |
| D2         | 1   | 0   | 0   | 0   | 1   | 0   |
| D#2        | 1   | 0   | 0   | 0   | 1   | 1   |
| E2         | 1   | 0   | 0   | 1   | 0   | 0   |
| F2         | 1   | 0   | 0   | 1   | 0   | 1   |
| F#2        | 1   | 0   | 0   | 1   | 1   | 0   |
| G2         | 1   | 0   | 0   | 1   | 1   | 1   |
| G#2        | 1   | 0   | 1   | 0   | 0   | 0   |
| A2         | 1   | 0   | 1   | 0   | 0   | 1   |
| A#2        | 1   | 0   | 1   | 0   | 1   | 0   |
| B2         | 1   | 0   | 1   | 0   | 1   | 1   |
| C3         | 1   | 1   | 0   | 0   | 0   | 0   |

Table II. Pin-out of DIP socket.

| <u>Pin</u> | <u>Function</u> |
|------------|-----------------|
| 1          | GROUND          |
| 2          | PLUG-5          |
| 3          | "A"             |
| 4          | "B"             |
| 5          | "C"             |
| 6          | "D"             |
| 7          | "E"             |
| 8          | "F"             |
| 9          | PLUG-14         |
| 10         | PLUG-13         |
| 11         | PLUG-12         |
| 12         | PLUG-11         |
| 13         | PLUG-10         |
| 14         | PLUG-9          |
| 15         | PLUG-8          |
| 16         | GROUND          |

\* \* \* \* \*



## VOLTAGE-CONTROLLED OSCILLATORS

VCO's: The heart of our VCO section will be a reprint of part of a presentation from EN#75, and two main options will be offered. A simple "utility VCO" will also be offered, which might well be a good first VCO if you are starting with a VCO and have not done much construction before. Also offered are three IC based VCO's with the first two having special features, and the third being more standard.

| <u>OPTION</u> | <u>SOURCE</u> | <u>DESIGNER</u> | <u>APPROX COST</u> | <u>NOTES</u>    |
|---------------|---------------|-----------------|--------------------|-----------------|
| VCO-1         | EN#75         | Hutchins        | \$20               | ENS-76 Option 1 |
| VCO-2         | EN#75         | Hutchins        | \$15               | ENS-76 Option 2 |
| VCO-3         | EN#67         | Hutchins        | \$3 - \$6          | Utility VCO     |
| VCO-4         | EN#65         | Hall            | --                 | Through zero FM |
| VCO-5         | EN#69         | Hall            | --                 | AM-FM           |
| VCO-6         | EN#87         | Rossum          | \$18               | SSM-2030        |
| VCO-7         | EN#129        | Hutchins        | \$22               | Through-Zero FM |



# THE ENS-76 HOME-BUILT SYNTHESIZER SYSTEM - PART 7, VCO OPTIONS:

-by Bernie Hutchins, ELECTRONOTES

## INTRODUCTION

We have chosen for the ENS-76 system six different VCO options which we will number 1, 1a, 2, 3, 4, and 5. Each of these has a combination of features which we feel make it a useful option. However, many of the features of one design can easily be used in another of the circuits, so the builder should feel free to experiment around. In this introduction, we will be briefly describing the different options. We shall be giving the main design features in the options, and for the benefit of those readers who are familiar with our earlier VCO designs, will point out the ideas that are new. This will permit the experienced designer to briefly scan the designs for new features without going over the circuit descriptions in detail.

VCO OPTION 1: This VCO is pretty much a standard VCO, with the exception that it has a Linear FM input which has been added because it is useful and easy to implement. The basic oscillator was given by Terry Mikulic in EN#62 (13). This option is intended to give the greatest accuracy and range of any of the designs. We used a somewhat simpler exponential stage here. The heart of the design is really the Analog Devices type AD818 NPN matched pair which has been optimized by the manufacturer for log conformance. A small amount of reset compensation (R\*) has been added to the design, but basically, we rely on the matched pair to keep the high end up (see the full circuit description for more information). The circuit has  $\pm 10$  volt signals throughout the waveshaping circuit, and it is trivial to bring out these signals, although we have chosen here to bring out our standard  $\pm 5$  volt signals. We have used a "counter-glitch" correction in the saw-to-triangle converter (see the Utility VCO in EN#67), and because we have a  $\pm 10$  volt triangle available, it is convenient to use the FET type triangle-to-sine converter.

VCO OPTION 1a: This VCO is identical to Option 1 except here we use the exponential converter used by Terry Mikulic in his original design. The advantage here is that the CA3046 transistor array costs under a dollar, while the AD818 pair in Option 1 runs five dollars in single units.

VCO OPTION 2: Option 2 is pretty much a "workhorse" VCO, but it does have one new feature (Symmetrized Ramp Modulation - SRM) and a number of new circuit tricks are given. For a discussion of the Symmetrized ramp, see the discussion by Bill Hartmann in EN#67. This circuit uses the triangle-square type of basic oscillator rather than the sawtooth oscillator in the designs above. In previous VCO's using the CA3080 as a current reversing switch, we have had to use an extra inverter or current mirror to properly drive the CA3080 and at the same time, have higher voltages correspond to higher frequencies. It is much simpler to just reverse the two base connections in the standard converter (we should have thought of using this earlier!). The basic oscillator is similar to those we have used before, but be sure to chalk up another trick to the CA3080's bag - it makes a good Schmitt trigger. We got this idea from the application notes on the type CA3140 op-amp (see RCA File No. 957). The waveshaping circuitry is standard, except we have used Bill Hartmann's triangle-to-saw method and implemented the symmetrized ramp modulation he suggests.

VCO OPTION 3: Option 3 uses a basic oscillator similar to that of Option 2, but here we have enhanced the capability for Dynamic Depth Linear Frequency Modulation (DDLFM). Note that this VCO has an input for an envelope generator! This envelope (or other control) is fed to an internal VCA which controls the depth of modulation. The VCO accepts external modulation, but will also accept self-modulation (harmonic spectrum). We have not put any waveshapers in this one. Instead, a single "spectral density" control changes the waveform through use of self-modulation.

VCO OPTION 4: In Option 4, we carry the linear modulation feature one step further to obtain both dynamic depth and through-zero modulation possibilities. See the work

of Douglas Kraul in EN#62 for more information on this type of oscillator. Since this is really a special purpose oscillator, and because a wide variety of timbres are available through the use of self-modulation, we have not added any of the conventional waveshapers to the circuit.

VCO OPTION 5: Option 5 is another special type of VCO, and is used mainly by way of illustration. We are calling it a "low distortion sinewave VCO" but it might well be considered a precision sinewave waveshaper. Here we have added two series single-pole low-pass filters to the usual triangle-to-sine converter to reduce the harmonic content (which is on the order of 1-2% out of the converter) to a much lower value. Note that we could mirror other circuits off the main converter in a similar manner. For example, we could have two more tracking oscillators for a "chord producing" VCO.

## GENERAL FEATURES OF ALL THE OPTIONS

All the VCO's have the standard exponential response of 1 volt/octave with control voltages fed in through 100k. The output levels are  $\pm 5$  volts, although it is generally possible to convert to a  $\pm 10$  volt system with little trouble. For the most part, signals are handled with the type 556 op-amp. In 5 volt systems, it is possible to use the 307 type op-amps if optimum high frequency performance is not required.

In order to simplify the schematic diagrams, we have not shown specific input stages but instead will be using as an input for control voltages the structure shown in Fig. 1. This could of course be used as is, but we have in mind that in nearly all cases the builder will use something with more features. In particular, COARSE and FINE frequency controls are common, and a variable input may also be desired. There should also be at least one input that is the direct-in 100k 1% type. A possible input stage is shown in Fig. 2.

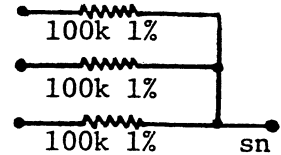


FIG. 1 INPUTS AS SHOWN

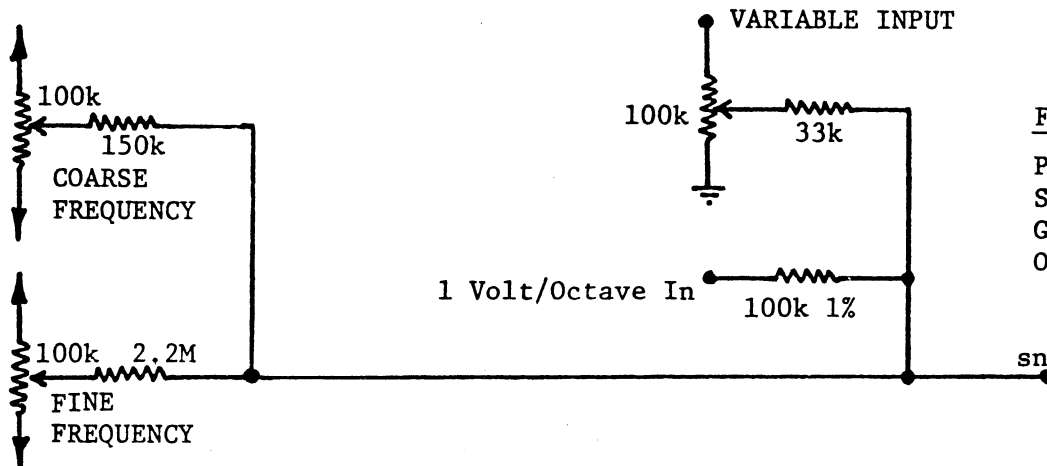


FIG. 2

POSSIBLE INPUT  
STAGE TO REPLACE  
GENERAL STRUCTURE  
OF FIG. 1

Beyond the summing node (sn) of the input structure, we will always find a scaling network which changes the 1 volt/octave at the input to the approximately 18 mV/octave that must appear at the base of a transistor. We also do a temperature correction at this point by means of the 2k T.C. resistor which is a Tel Labs type Q81k which has a positive T.C. of +3500 ppm/°C. This corrects for temperature variations in the transistors of the exponential converter. Thus, both this resistor and the transistors should be at the same temperature, and this is indicated by the "thermal contact" arrows. A typical input is shown in Fig. 3:

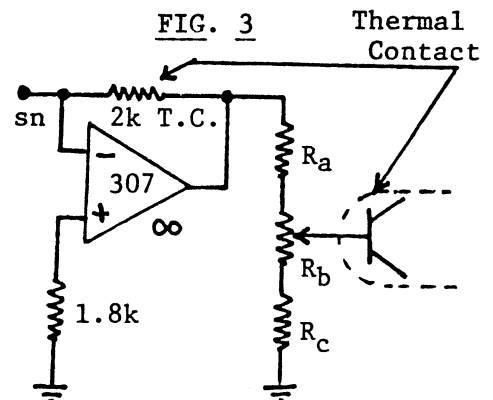


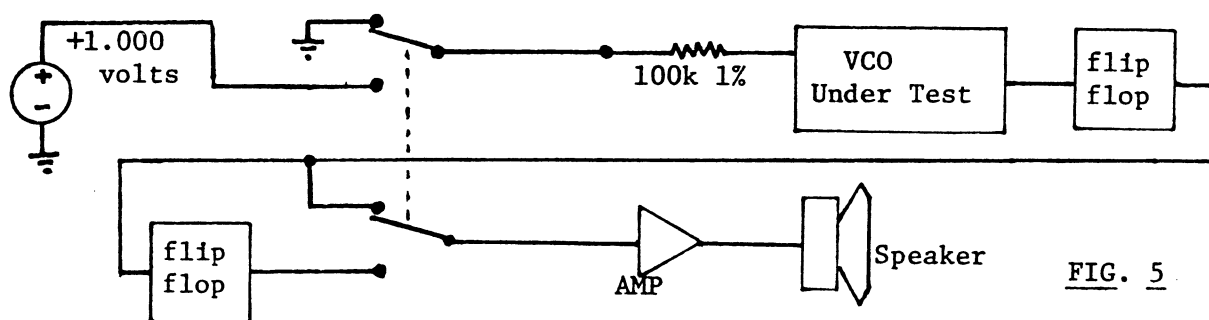
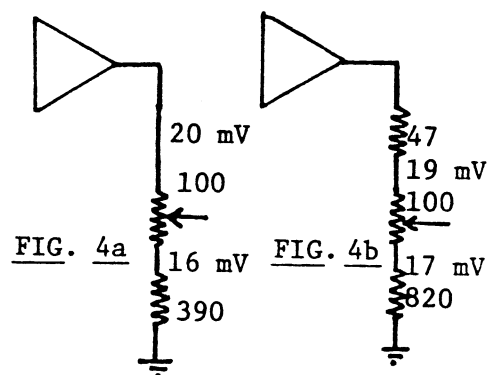
FIG. 3



In practice, thermal contact is made by mounting the resistor and the transistor pair or array so that they are touching, and then putting a small amount of heat-sink compound between them. For set-up purposes and testing, an ordinary 2k 5% resistor can be used. The T.C. resistor and heat sink compound can be put in during final packaging and calibration.

The resistors  $R_a$ ,  $R_b$  (a trim pot, probably a multi-turn unit), and  $R_c$  are selected so that a one volt change at the input, through 100k, is scaled to an 18 mV change at the wiper of the trim pot. At the same time, it is best to maintain a low source impedance (say 1k or lower) so that the base of the converting transistor is held down well. Since the output of the op-amp will never exceed a voltage of much more than a few hundred mV, there is no problem with the op-amp driving such a low impedance. In the circuit diagrams, we will be showing the voltage divider shown in Fig. 4a ( $R_a = 0$ ,  $R_b = 100$ ,  $R_c = 390$ ) which permits the voltage to be scaled from any values between 20 mV and 16 mV. This works quite well if a multi-turn pot is used, and gives a lot of excess room on the ends. A somewhat tighter divider is shown in Fig. 4b where  $R_a = 47$ ,  $R_b = 100$ , and  $R_c = 820$ . This gives voltages scaled from 17 mV to 19 mV. If you are using the precision resistors indicated (the 100k 1% or better), this should work out well. If however you are just setting up with 5% resistors, you may need the extra room given by the circuit of Fig. 4a.

In any case, you will be adjusting the trim pot so that a 1 volt change at the input gives a one octave change of frequency. This should be set first at the lower frequencies (around 100 Hz) and then checked at higher frequencies, and compensated if necessary. Tuning is often best done by ear. As an aid, it is useful to have a precision one volt source which is switched in, and at the same time, a flip-flop is switched into the return line from the VCO. Thus, the pitch of the VCO should go up one octave, but the pitch returning should remain the same, because of the additional flip-flop. The original flip-flop is needed so that only square waves are compared. See Fig. 5 for this setup.



Most of the other adjustments and calibrations that you will need to make are described in the circuit descriptions of the various options. Keep in mind that there are six different circuits, and we built each one only once. If you copy ours exactly, it should work, but you may have to do a little trimming to get everything exactly the way you want it. We were not overly careful to adjust the amplitudes to exactly the five volt levels, and made no trim adjustment provisions for amplitude except in one case. You should trim these up if you find it necessary. You can use a scope to do this if you have an accurate one. However, even if you don't have an accurate scope, you can use an accurate meter. You just set the VCO frequency to 1/10 Hz or below and read the peak of the meter. For making adjustments of waveshapers, it is best to use a scope, even if it is a relatively poor one. Note however, that when trimming up the sine wave, the best instrument you have is your ear.

## ENS-76 VCO OPTION 1

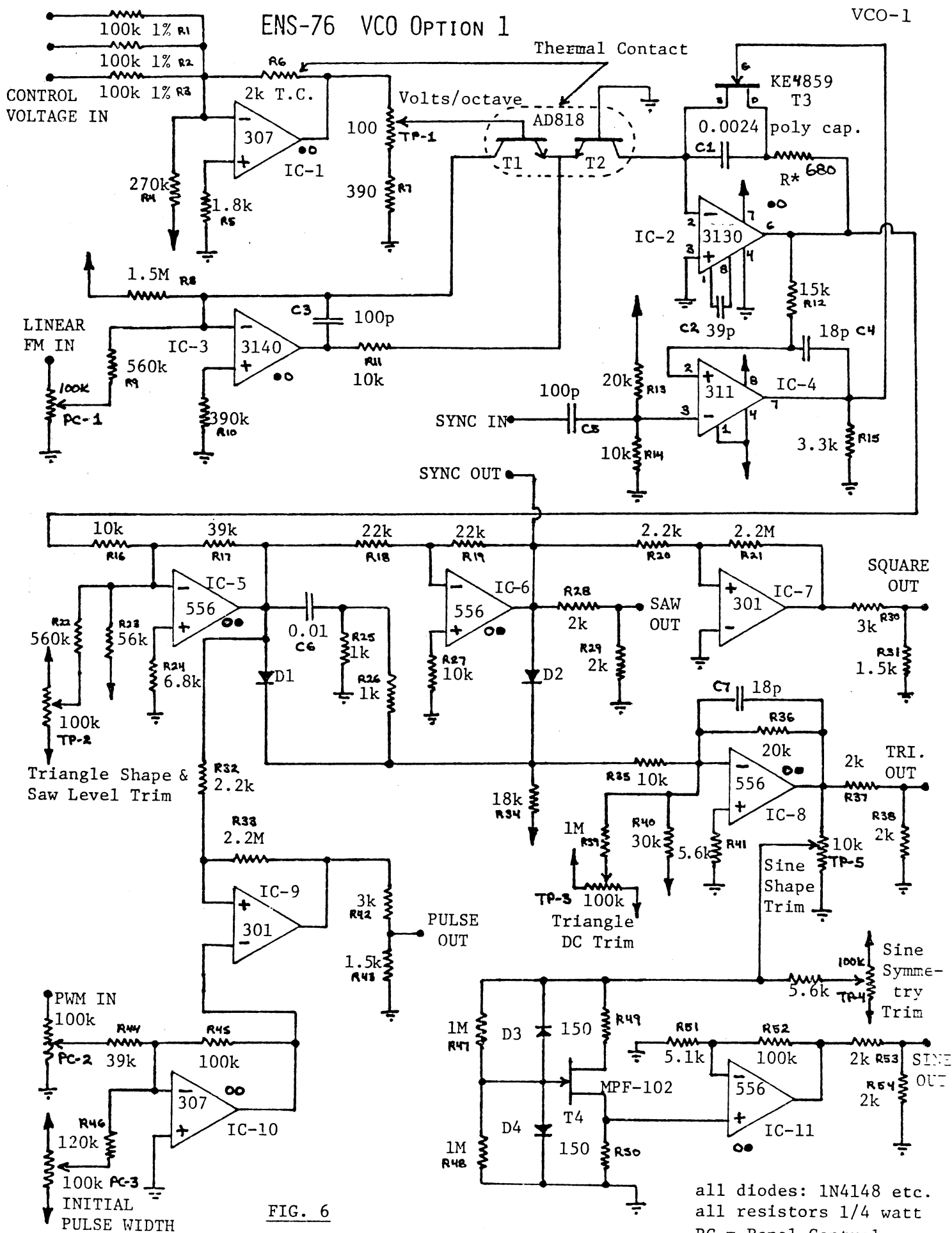
The design goal for VCO Option 1 was to make a VCO with standard features that is as accurate and drift free as possible. The basic oscillator (IC-2 and IC-4) is a sawtooth generator based on the VCO described by Terry Mikulic in EN#62. This basic oscillator is driven by an exponential current source based upon the Analog Devices type AD818 matched NPN pair, which is probably the best exponential (log) transistor pair available. The output of the basic oscillator is a zero to +5 volt ramp (output pin 6 of IC-2). The rest of the circuit consists of waveshapers that are more or less standard, and provide Saw, Triangle, Square, Pulse, and Sine outputs, all with  $\pm 5$  volt levels. It is easy to convert this one to  $\pm 10$  volt levels if this is desired. The circuit also has inputs for Linear Frequency Modulation, and for standard Pulse Width Modulation.

The exponential current source (IC-1, IC-3, T1 & T2) is a standard design which we have used many times before. Here we have improved things as much as possible by using the CA3140 op-amp to regulate the reference current, and the AD818 pair as the converting transistors. No provisions are made here for high end compensation of the exponential stage, but standard temperature compensation is made.

The basic oscillator is a sawtooth (ramp) relaxation oscillator formed from IC-2 and IC-4, and is only slightly changed from the original Mikulic design. IC-2 forms an integrator where the exponential current from the collector of T2 causes capacitor C1 to charge. For the moment, assume that  $R^*$  is zero ohms. When the output of IC-2 reaches +5, the voltage on pin 2 of IC-4 is the same as the reference on pin 3 (as determined by voltage divider R13-R14), and as the voltage starts to rise above +5, comparator IC-4 goes from -15 to ground (pulled up by R15). This turns on the FET switch T3, and capacitor C1 begins to discharge rapidly. In general, the capacitor would discharge a little below +5 and the comparator would then shut off the FET switch. This would result in an output consisting of a high frequency oscillation of low amplitude centered about +5. This is where the capacitor C4 comes in. When the comparator switches high (to ground that is), the capacitor voltage is +20 volts (-15 from the comparator, and +5 fed in from the ramp through R12). This will cause the comparator to remain high until C4 discharges to +5 volts (starting at +20 and being discharged through R12 to pin 6 of IC-2, which is going from +5 to zero). The time constant is thus something like  $1.4 \cdot R12 \cdot C4$ , which is something like 400 ns. This is the time interval during which the FET switch will be on. This is long enough to completely discharge the capacitor for all practical purposes. Note finally that the discharge can be initiated a little early by the Sync. control input. A negative going pulse here will lower the reference a bit and institute the reset cycle.

The sawtooth-to-triangle converter consists of IC-5, IC-6, and IC-8. IC-5 takes the zero to +5 ramp from IC-2, amplifies it by a factor of 4 and level shifts it. The result is an inverted sawtooth from +10 to -10. IC-6 is an inverter which gives back the normal (rising) sawtooth from -10 to +10. This is brought out as the sawtooth output, attenuated and impedance adjusted by R28-R29 to a ramp from -5 to +5 with an output impedance of 1k. The diode arrangement of D1 and D2 selects the higher of the two waveforms - the sawtooth or the inverted sawtooth. This is easily seen to be a triangle from +10 to zero. IC-8 amplifies this by a factor of two and level shifts for zero DC average. The output of IC-8 is thus a +10 to -10 triangle. The output is obtained through R37 - R38 as a  $\pm 5$  triangle. We have used a "counterglitch" circuit here to remove (to a large degree) the switching glitch caused by the diodes. This counterglitch is provided from the inverted sawtooth through C6 and R26. Some minor adjustments of the value of C6 may be desirable in some setups. R34 pulls down the diodes, resulting in a better positive "point" on the triangle output, and C7 acts to further remove any switching glitches.

Rectangular waveforms are provided using uncompensated 301 type op-amps (or 748 types can be used). The square is provided by IC-7, and a variable pulse is provided



All Diodes: 1N4148 EN#75 (7)

all diodes: 1N4148 etc.  
all resistors 1/4 watt  
PC = Panel Control  
TP = Internal Trimmer

by IC-9, relative to a reference voltage level provided by IC-10. This is all very standard, but keep in mind that both IC-7 and IC-9 are left completely uncompensated so that they will go as fast as possible (no capacitor between pins 1 and 8), and the positive feedback provided by the 2.2M resistors relative to the 2.2k input provides mild hysteresis for additional "snap" and noise rejection. The outputs of IC-7 and IC-9 range from +15 to -15, so 3k-1.5k voltage dividers are used to scale these to  $\pm 5$  volt levels.

Finally, we use the FET type sinewave shaper as driven by the triangle waveform. We chose this rather than the CA3080 shaper (used in later options) because it gives a slightly better looking waveform, and because we do have the  $\pm 10$  volt triangle to drive it. It normally requires a 6 to 7 volt signal to reach the non-linear region.

It is probably obvious that since we are using  $\pm 10$  volt levels inside the circuit, it is easy to bring these out. For easy reference, the conversion is listed below:

| <u>RESISTOR</u> | <u>AS IS <math>\pm 5</math></u> | <u>CONVERSION TO <math>\pm 10</math></u> |
|-----------------|---------------------------------|------------------------------------------|
| R28             | 2k                              | 1k                                       |
| R29             | 2k                              | Omit                                     |
| R30             | 3k                              | 1.5k                                     |
| R31             | 1.5k                            | 3k                                       |
| R37             | 2k                              | 1k                                       |
| R38             | 2k                              | Omit                                     |
| R42             | 3k                              | 1.5k                                     |
| R43             | 1.5k                            | 3k                                       |
| R53             | 2k                              | 1k                                       |
| R54             | 2k                              | Omit                                     |

At this point, we want to say something about the resistor  $R^*$ , which is the resistor that provides high frequency compensation by Franco's method [See S. Franco, "Hardware Design of a Real-Time Musical System", Dept of Computer Science Report UIUCDCS-R-74-677, Univ. of Illinois, Urbana-Champaign]. The principle is basically as follows: There is a finite switching time which causes an oscillator to go flat on the high end. The higher frequencies correspond to higher charging currents. Thus, by inserting a resistor in series with the integrating capacitor, an additional voltage is impressed across the R-C combination, and this causes the oscillator to reach its peak voltage a little earlier. A full analysis (see reference above or chapter on VCO's in MEH) will show that this is an exact correction for a constant reset time, and that the  $R^* \cdot C_1$  product should be equal to the switching time, which we saw above was about 400 ns. This gives a value of  $R^*$  of about 166 ohms. This is a good starting value. In addition, it turns out that the error due to the bulk base-emitter resistance of the exponential current stage transistors can be corrected by a term that is the same order as the delay time correction. Thus, by making  $R^*$  larger than is necessary to correct for delay time, we can also do some correction for bulk resistance. Of course, all the analysis in the world is generally no substitute for an actual experiment, and we can easily find the value for  $R^*$  by an experiment - basically as shown in Fig. 5.

The data from an actual experiment is shown in Fig. 7. Here we have tabulated control voltage, approximate frequencies, and the ratio of frequencies that we actually observed for a one volt change of control voltage. Ideally, this should be 2.00 in all cases. By studying these figures, you can see that 680 ohms of Franco compensation pretty much solves the problem over the audio range. It is well not to use much more resistance than this as the method does result in some imperfections in the waveforms, and the oscillator may stall at low frequencies. Note however that the imperfections that result from the 680 ohm resistor are not too important as they only become appreciable at frequencies above 5 kHz or so, and the harmonic content in waveforms at such frequencies is of little importance because much of it is beyond the upper frequency limit of the ear.

|        | <u>CONTROL VOLTAGE</u> | <u>APPROX FREQ.</u> | <u>UNCOMPENSATED</u> | <u>680 OHM FRANCO COMP.</u> |
|--------|------------------------|---------------------|----------------------|-----------------------------|
| FIG. 7 | -1.00                  | 10                  | 2.00                 | 2.00                        |
|        | 0.00                   | 20                  | 2.00                 | 2.00                        |
|        | +1.00                  | 40                  | 2.00                 | 2.00                        |
|        | +2.00                  | 80                  | 2.00                 | 2.00                        |
|        | +3.00                  | 160                 | 2.00                 | 2.00                        |
|        | +4.00                  | 320                 | 2.00                 | 2.01                        |
|        | +5.00                  | 640                 | 2.00                 | 2.00                        |
|        | +6.00                  | 1280                | 1.99                 | 2.00                        |
|        | +7.00                  | 2560                | 1.99                 | 2.00                        |
|        | +8.00                  | 5120                | 1.97                 | 2.00                        |
|        | +9.00                  | 10240               | 1.93                 | 1.97                        |
|        | +10.00                 | 20480               |                      |                             |

It should not be too difficult to construct this circuit. A little care should be used handling the MOS op-amps (CA3130 and CA3140), and of course be careful with the expensive AD818 pair. All op-amps are powered between +15 and -15 except for IC-2, the CA3130. Also, be sure to note the unusual pin connections for the LM311 comparator, and be sure not to forget the pull-up resistor R15. All the type 4859 FET's we have seen have the base diagram shown in Fig. 8, as seen from the bottom view. This includes the 2N4859, the KE4859, and the PN4859. Some may have three in-line wires at the very base, but will have the triangular diagram further out.

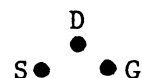
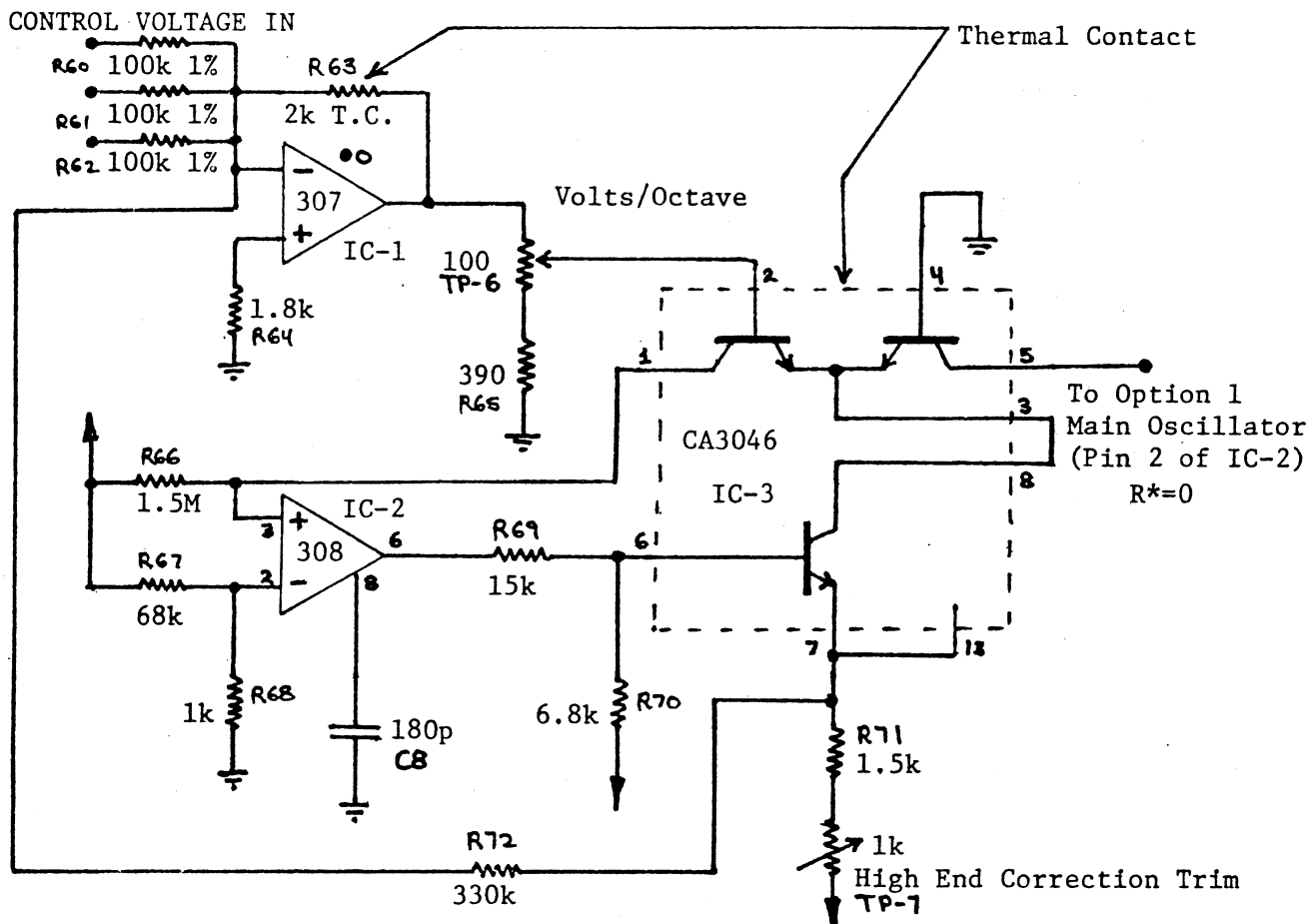


FIG. 8

## ENS-76 VCO OPTION 1A

Option 1a is a variation on Option 1. In 1a we are using the original exponential current source given by Terry Mikulic in EN#62 which uses Moog-Hemsath compensation rather than Franco compensation. Thus, we have only to use the exponential stage of Fig. 9 connected to the oscillator and waveshapers of Fig. 6, and make  $R^*=0$ . This circuit also works quite well. The principal advantage is the lower cost of the CA3046 array as compared to the AD818 pair, and the principal disadvantage is that it has one additional trim pot to tune up. Some builders may want to leave in  $R^* = 150$  ohms for Franco compensation of the reset time, and use the correction shown here for only the bulk resistance correction. We have not looked at this carefully to see which is the better way.

Be sure to note that IC-2 is an LM308 type op-amp. We suggest only this one here. A 307 with no external compensation may work with a little oscillation at the output, but a CA3140 does not work, and placing a capacitor between pins 3 and 6 of the CA3140 will of course cause additional oscillation rather than compensate. There may be a way of compensating with the CA3140, but unless you love experimenting, stick with the LM308 here with the 180 pf compensation shown. Also, don't forget to connect up the substrate pin (pin 13) of the CA3046 as shown, or connect it through about 10k to the -15 supply.



To adjust the tracking of Option 1a, you first set the high end correction trimmer to its minimum value and tune the 100 ohm volts/octave control for 1 volt/octave in the low frequency (say 100 Hz to 200 Hz) range. Next, adjust the input control voltage so that you are in the upper frequency range (say the 7500 Hz to 15,000 Hz octave) and check the ratio here. It should be less than 2.00. Now, adjust the 1k High End correction pot in the direction so that the series resistance of the pot and R71 increases. This will increase the feedback. Once you get a one volt/octave response up high, recheck the lower octaves. There should be no change, and the overall tracking should be quite good.

## ENS-76 VCO OPTION 2

VCO Option 2 is shown in Fig. 10. This option, like Option 1 and Option 1a above is a real "workhorse" option. Even if you decide to make mostly Option 1 oscillators, you should have at least one Option 2 as it has several additional features of interest.

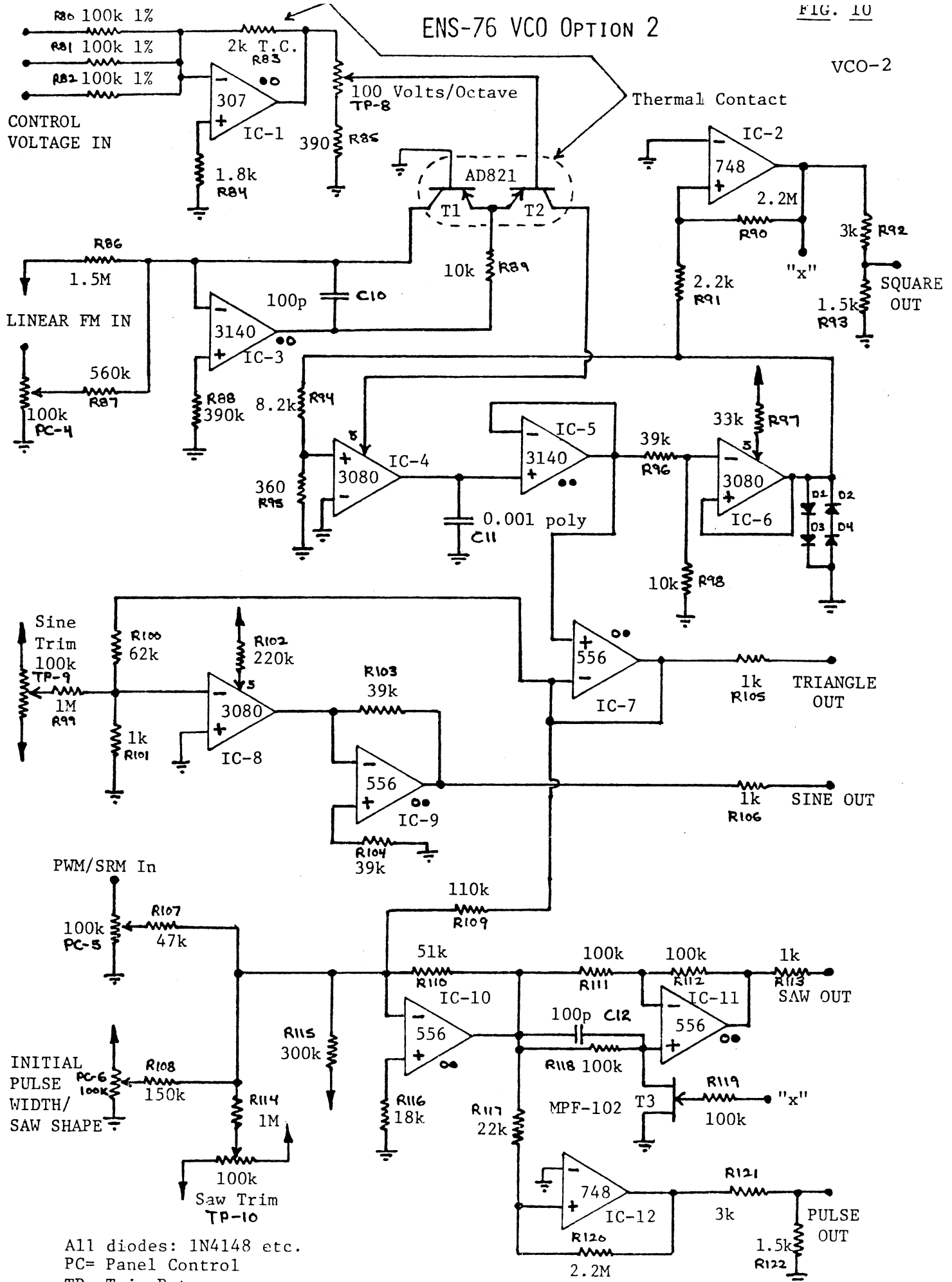
Basically, Option 2 consists of an exponential current source IC-1, T1 and T2, IC-3; Triangle-Square oscillator IC-4, IC-5, IC-6; and associated waveshapers.

We have used the CA3080 as a current reversing switch before (EN#46 for example). The CA3080 must be driven by a PNP transistor, and hence, we have had to use the AD821 matched PNP pair, which works just fine. However, we have also had to add an extra inverter or current mirror to get higher voltages to correspond to higher frequencies, which is the usual case. These extra elements of course add to the possibility of additional drift. A much simpler solution is to just reverse the base connections of

# ENS-76 VCO OPTION 2

FIG. 10

VCO-2



All diodes: 1N4148 etc.  
PC= Panel Control  
TP= Trim Pot

the two exponential converter transistors. Note that this means that the control voltage (as scaled to about 18 mV/Oct.) is applied to the base of the same transistor that supplies the actual exponential current. Compare this with Option 1 in Fig. 6 which uses an NPN pair. This simplification can be applied to other circuits we have given as well.

The basic oscillator consists of IC-4, a current reversing switch driving integrating capacitor C11; a FET input op-amp voltage follower IC-5, and a fast Schmitt trigger. Note that diodes D1 to D4 limit the output of IC-6 to either about +1 or -1, and that this voltage is fed back to the + input of IC-6. Suppose first that the + input of IC-4 is positive, driven by the output of IC-6. This means that the control current into pin 5 of IC-4 is being driven into C11. Thus the voltage across C11 is increasing in a positive direction, and IC-5 is following this voltage. When the output of IC-5 reaches about +5 volts, voltage divider R96-R98 will be supplying about +1 volts to the - input of IC-6, and the output of IC-6 will go to -1 volts. This means in turn that the capacitor voltage will start to ramp downward. The cycle repeats producing a triangle wave of 5 volt amplitude.

We have used this type of basic oscillator before, but have not used the CA3080 as a Schmitt trigger. In fact, it appears to be a good one and is plenty fast for an audio frequency oscillator. One possible drawback is the fact that the diodes D1 to D4 determine the amplitude of the triangle, and thus indirectly the frequency. Since silicon diodes exhibit approximately a  $-1.7 \text{ mV}/^{\circ}\text{C}$  change of forward voltage drop, this is of some concern, because this is about  $.28\%/^{\circ}\text{C}$ . In a rough experiment, we measured a drift of only about 1/3 this amount for the entire oscillator. It would seem that this deserves more study, but there may be no real problem.

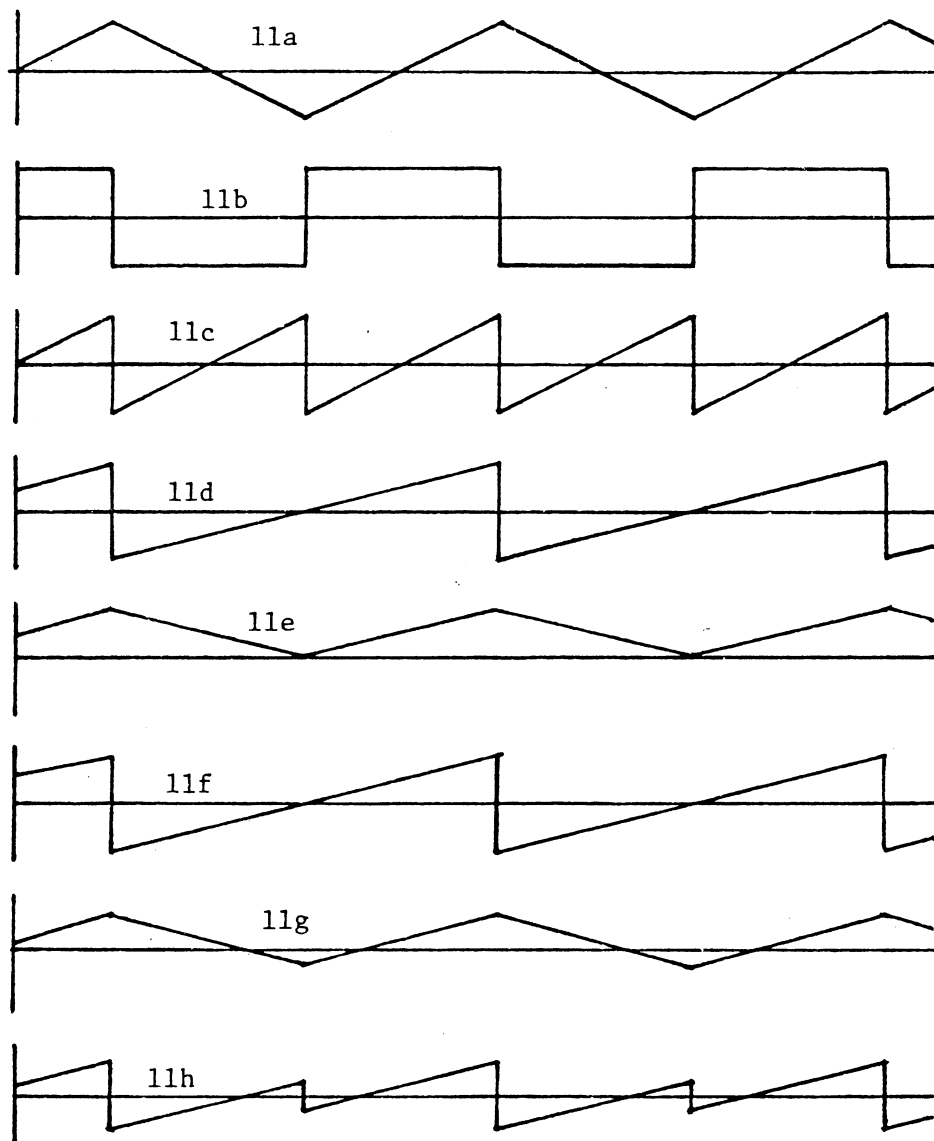
Since the CA3140 is not suitable for driving 1k loads in our application (see EN#69, pg. 13), we buffer the triangle with a 556 before bringing it out. You could use the 556 as IC-5, omitting IC-7, but the CA3140 is excellent as the follower on the capacitor due to its low bias current requirements.

IC-8 and IC-9 form a triangle-to-sine converter using the overdriven differential amplifier principle (see MEH, Chapter 5b (18)). The sine trim is adjusted for symmetry of the upper and lower portions of the waveform. By ear, one just adjusts for lowest distortion. In fact, the adjustment shown removes only even harmonics, but the 62k-1k voltage divider pretty well optimizes the driving voltage for lowest odd harmonic distortion. Note also that a square wave is obtained using IC-2 in a conventional manner.

It remains to bring out the pulse, and to convert the triangle to a sawtooth. We will do these at the same time as we want our Pulse Width Modulation (PWM) setup to be used also in a Symmetrized Ramp Modulation (SRM) setup. The principles involved are illustrated in Fig. 11. A discussion of the original idea was given by Bill Hartmann in EN#67. Fig. 11a shows the original triangle output while Fig. 11b shows the square wave "x" in the circuit. If we consider that the square can be made to invert the triangle wave if it (the square) is negative, and to not invert if it (the square) is positive, we can produce the double frequency sawtooth of Fig. 11c using a circuit like that of IC-11 and surrounding components in Fig. 10. We did this in our older oscillators, and then added a portion of the square wave to get the original frequency back (see Fig. 11d). Bill Hartmann suggested that instead we first add a DC voltage to the triangle so that it touches zero, but is otherwise always positive (Fig. 11e) and then invert-or-not-invert the waveform using the square. This gives the original frequency sawtooth as before (Fig. 11f). Now, by allowing the DC offset to take on a general level (Fig. 11g for example), a Symmetrized Ramp results (Fig. 11h). By varying the DC level, we get modulation of the symmetrized ramp. This is exactly what we do in the Option 2 circuit. The offset is provided by IC-10. It is then a trivial matter to obtain Pulse Width Modulation (PWM) by zero crossing the offset triangle (using IC-12).



FIG. 11



It is probably easiest to see how Symmetrized Ramp Modulation works by studying Fig. 12. In Option 2, both the SRM and PWM processes are driven by the offset triangle. The DC offset of the triangle is the modulation voltage. In Fig. 12, the triangle is shown with five different DC levels. Above the offset triangles is shown the square wave (point "x" of Fig. 10) which determines whether the offset triangle is output in its normal or inverted form. A full swing of the DC offset changes the waveform from a normal rising sawtooth, through a double frequency sawtooth, back to another normal sawtooth 180° out of phase with the original.

The pulse waveform is quite easy to understand, as it is just a zero-crossed version of the triangle. As we have shown it, there is no output from the pulse output when the triangle is fully positive, but there is a very narrow pulse out when the triangle is a little negative. This is really determined by the way the "Saw Trim" pot is set. Actually, in our setup this was the way it came out when we set the best saw shape.

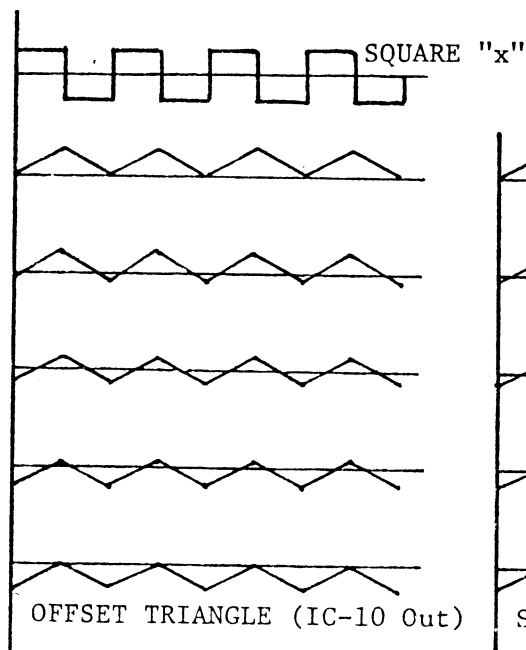
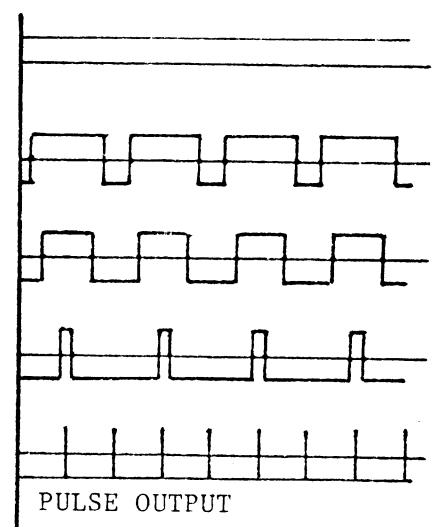
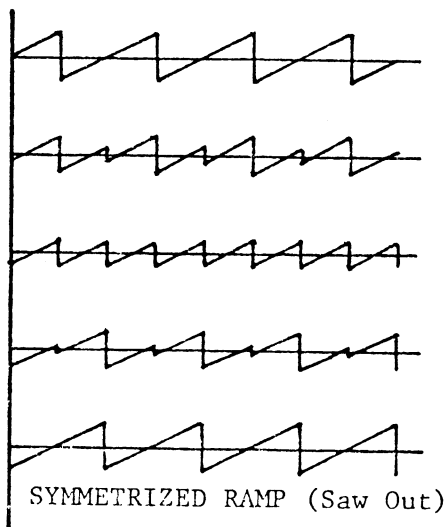


FIG. 12



If you prefer, you can balance things out so that you have no pulses out at all when the normal sawtooth is set, or so that you have narrow pulses out for both normal sawtooth positions. As we have set it up, you have a choice. In one normal sawtooth position, you can get a DC output of +5 volts from the Pulse Out. In the other normal sawtooth position, you get narrow pulses. The feeling is that in many cases, the players will be using mostly normal waveforms, and it may be confusing if there is no output from the pulse at the time when the sawtooth is quite normal. What it boils down to is that you have a choice as to which extreme of the pot rotation you choose to mark as normal sawtooth. Note that we have added a little extra hysteresis to this pulses circuit (we made R117 = 22k rather than 2.2k) because of the closeness of the offset triangle to ground. By the way, R109 really is 110k in our setup, not 100k.

The volts/octave control is set up in the normal manner: first you adjust a lower octave and then check the higher octaves. In our setup, we actually found that the high-frequency performance went up, rather than down as you would normally expect. It would probably not be too difficult to slow this down if it is necessary. Note in particular that over the normal audio range, the performance is quite good. The data table is given in Fig. 13.

Note that for the normal sawtooth, there is a glitch in the midpoint of the sawtooth due to switching. We found that the 100pf capacitor C12 did a good job of greatly reducing this glitch. The builder may want to play around with this value a little.

A number of additional changes to the circuit are possible. For example, if you add a zero-cross detector to the triangle out of

IC-5, you can use this to drive point "x" instead of the output of IC-2. A switched option would be suggested if this is done. The result is a whole new group of waveforms is you can easily show with a few sketches, since the new square wave is out of phase with the square wave out of IC-2 by 90°. Also, by inverting the square wave out of IC-2, you get inverted rather than normal sawtooth waveforms. It is also interesting to form mixtures of pulses and symmetrized ramps, and you may want to make this a standard feature.

| CONTROL VOLTAGE | APPROX. FREQ. | RATIO |
|-----------------|---------------|-------|
| -8.00           | 1.5           | 2.00  |
| -7.00           | 3.0           |       |
| -6.00           | 6.0           | 2.00  |
| -5.00           | 12.0          |       |
| -4.00           | 24.0          | 2.00  |
| -3.00           | 48.0          |       |
| -2.00           | 96.0          | 2.00  |
| -1.00           | 192           |       |
| 0.00            | 384           | 2.00  |
| +1.00           | 764           |       |
| +2.00           | 1528          | 2.01  |
| +3.00           | 3056          |       |
| +4.00           | 6112          | 2.01  |
| +5.00           | 12224         |       |
| +6.00           | 24448         | 2.01  |
| +7.00           | 48896         |       |

FIG. 13 Option 2 Data

## Oscillators VCO-3 through VCO-6

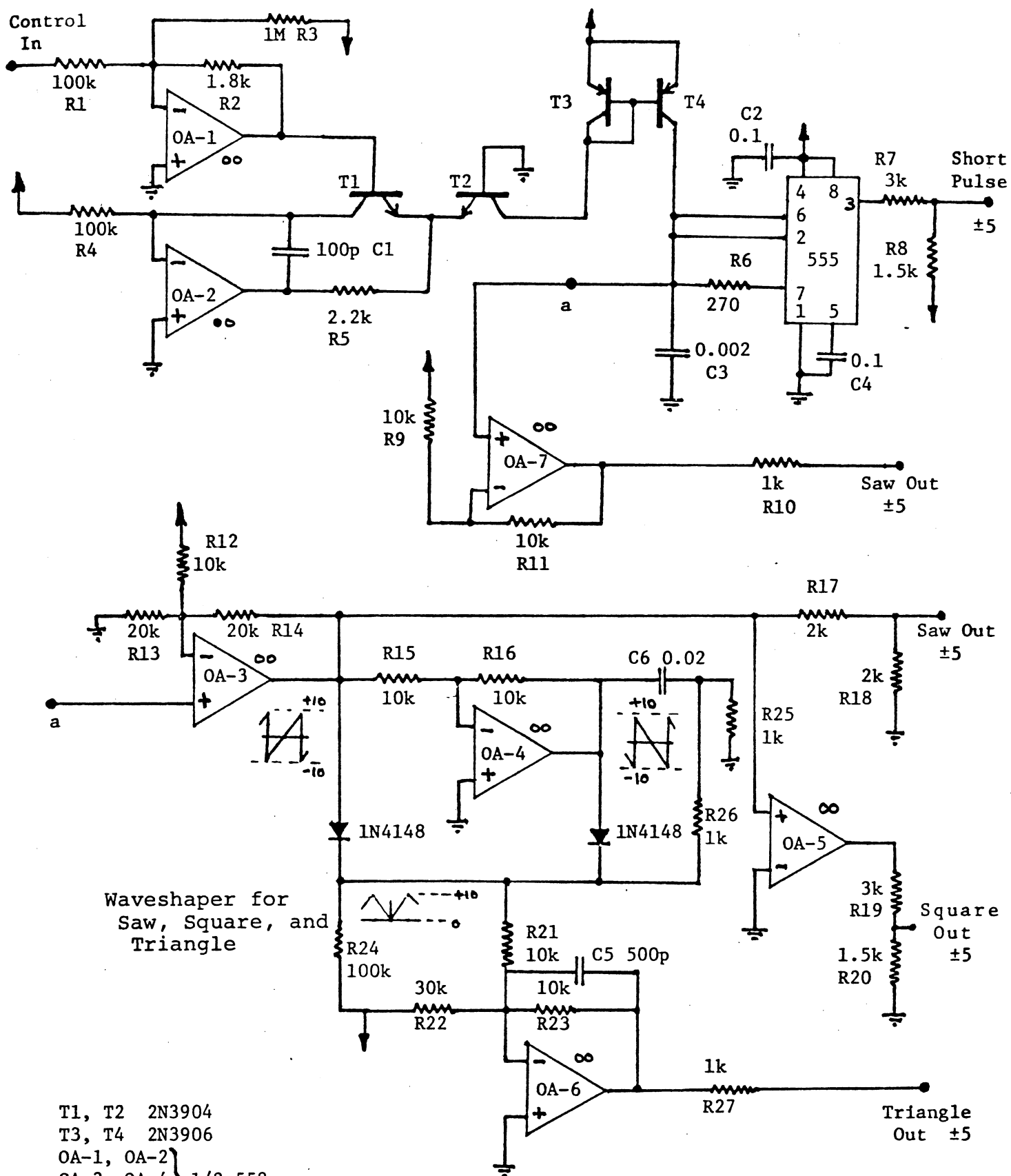
The last four VCO's will be described briefly. For a full discussion, see the description in the appropriate newsletter (see page number at the bottom of schematic).

VCO-3: Utility VCO. Note that the top of the circuit can be built alone and is a "dirt cheap" and simple oscillator.

VCO-4: A special purpose oscillator for "through zero" FM.

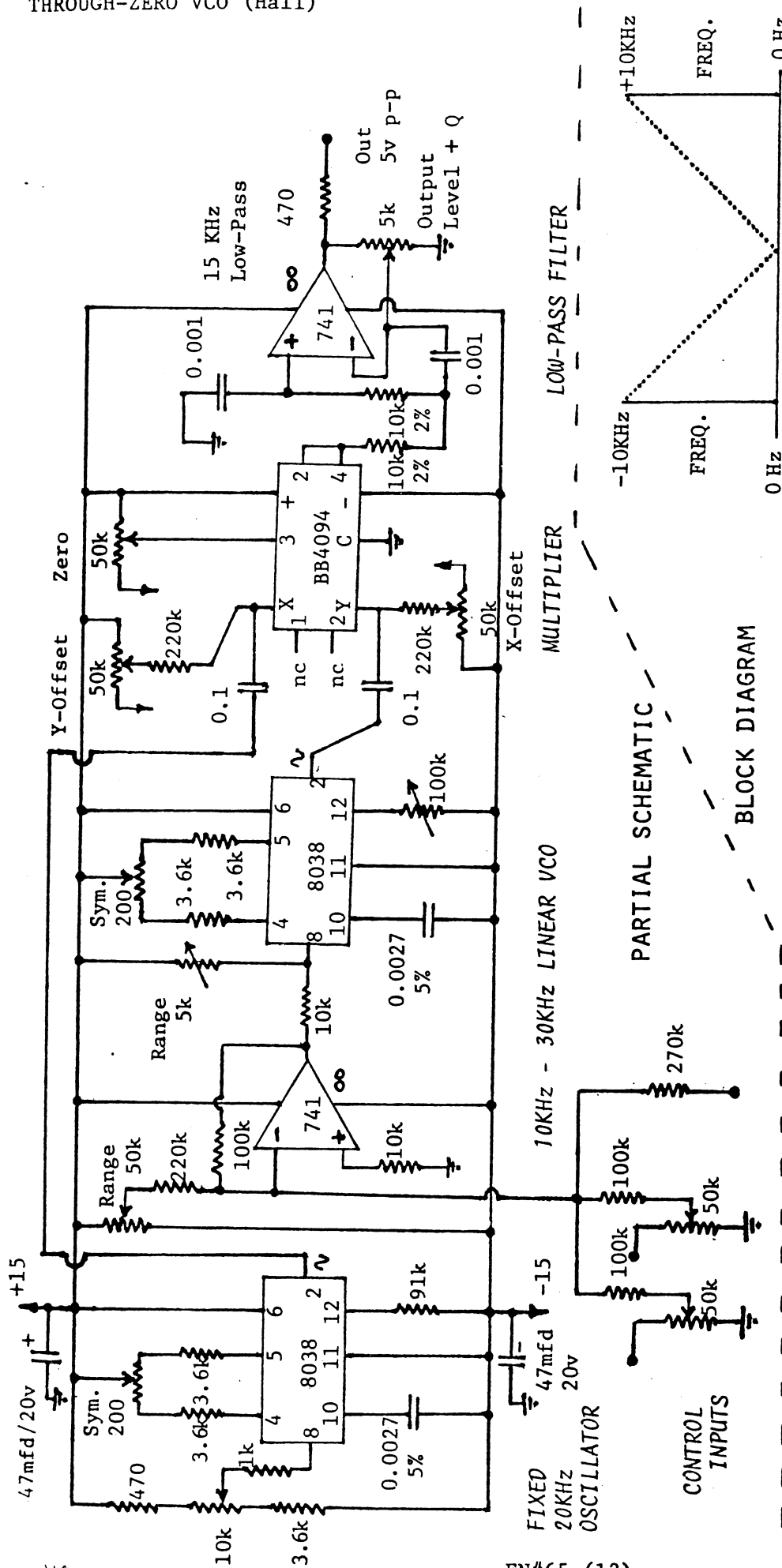
VCO-5: An oscillator taking advantage of an existing VCO IC Chip.

VCO-6: This VCO uses the SSM2030 VCO IC which is a unit made specially for electronic music work,

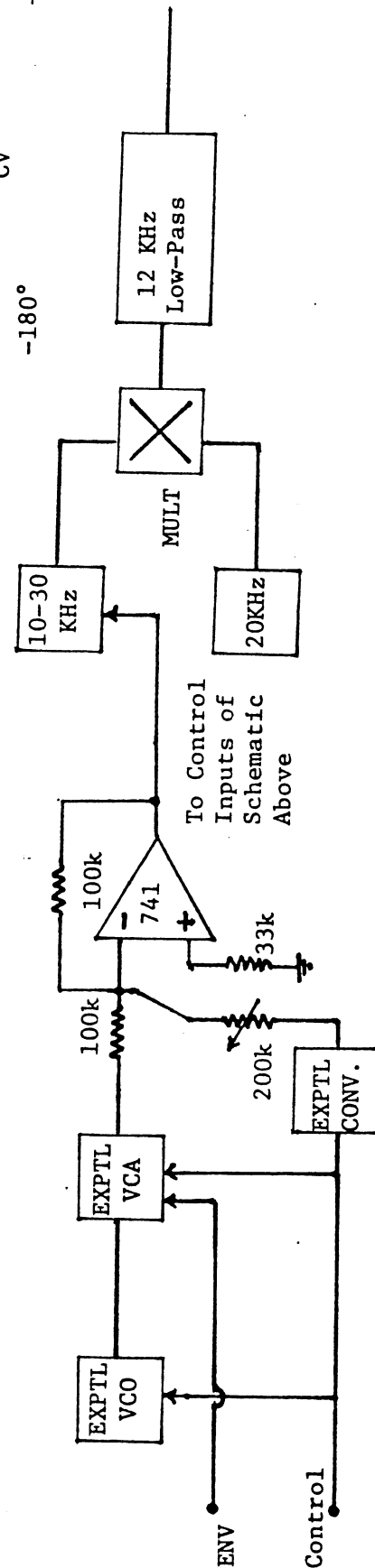


T1, T2 2N3904  
 T3, T4 2N3906  
 OA-1, OA-2 } 1/2-558  
 OA-3, OA-4 }  
 OA-5, OA-6 }  
 OA-7 307

ENS-76 "UTILITY VCO"



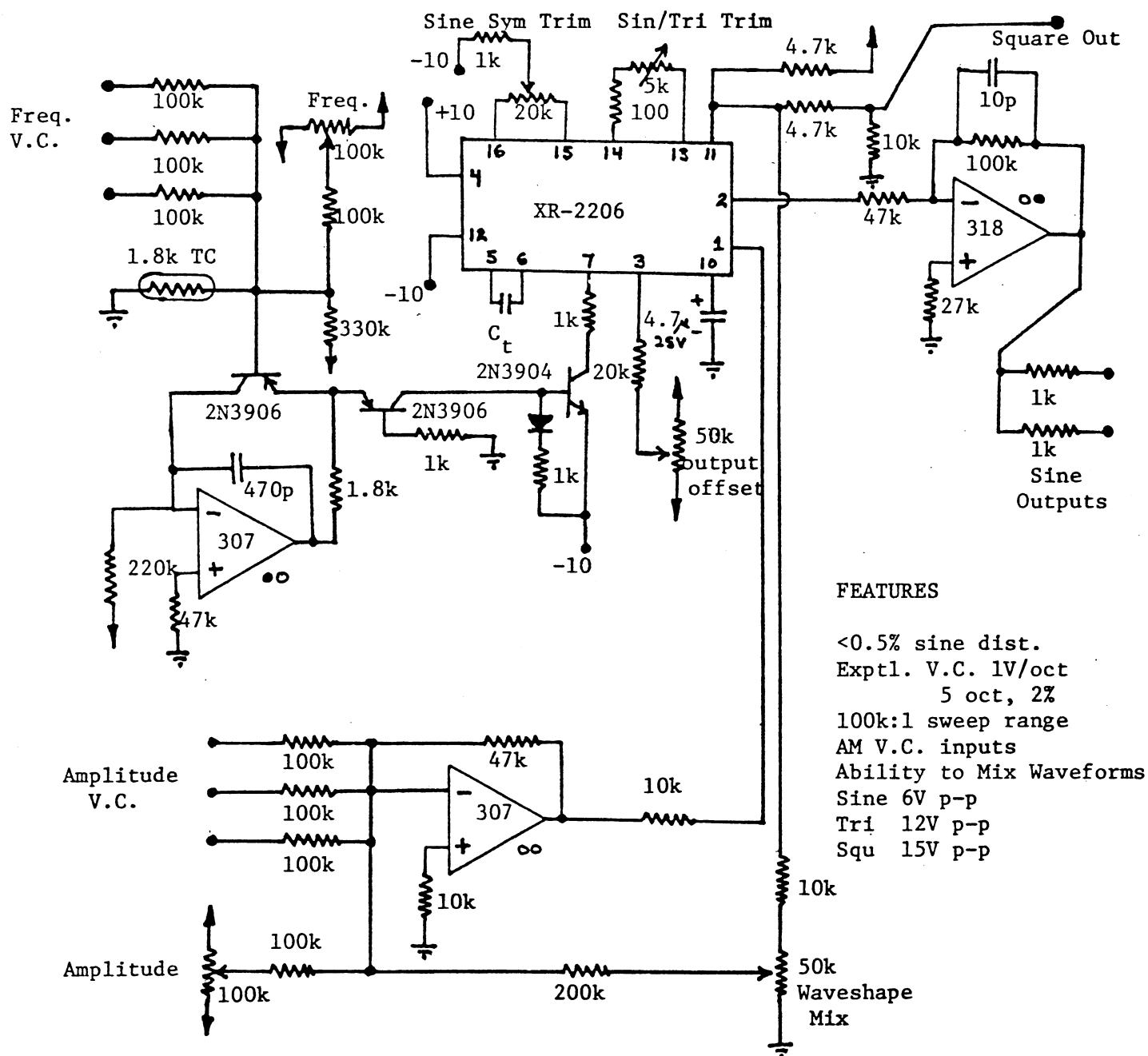
BLOCK DIAGRAM



Schematic Not Shown Above → Schematic Shown Above →

FIG. 1 AM-FM Utility VCO Using XR2206 VCO Chip

- J. R. Hall

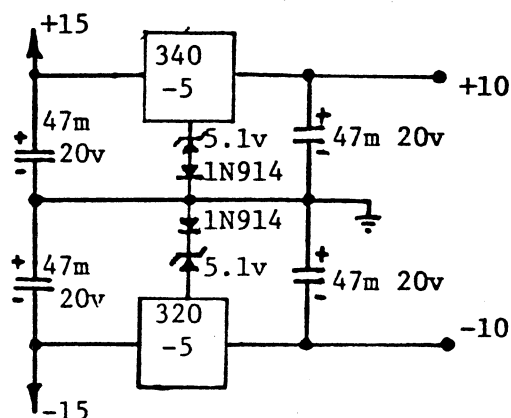


## FEATURES

- <0.5% sine dist.
- Exptl. V.C. 1V/oct
- 5 oct, 2%
- 100k:1 sweep range
- AM V.C. inputs
- Ability to Mix Waveforms
- Sine 6V p-p
- Tri 12V p-p
- Squ 15V p-p

FIG. 2

±10 Volt Regulator



Notes:  $C_t = 0.01$   
 0.5 Hz - 50kHz  
 $C_t = 1 \text{ mfd}$   
 0.01 Hz - 50 Hz

1.8k T.C., both  
 2N3906's, and the  
 2N3904 should be in  
 close proximity



# A VOLTAGE-CONTROLLED OSCILLATOR WITH THROUGH-ZERO FM CAPABILITY:

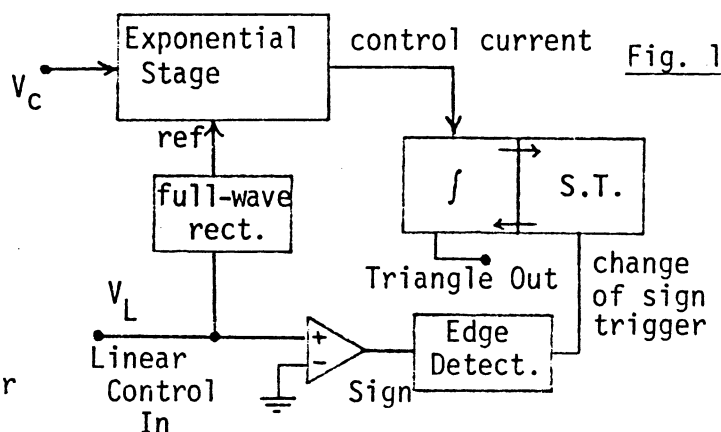
-by Bernie Hutchins

## INTRODUCTION:

It seems that musical engineers just can't go for long periods of time without trying out new VCO designs. Indeed much space in earlier issues was devoted to the art of designing VCO's, each one adding new tricks. Much of the "fun" of designing VCO's has been taken from us with the appearance of single-chip, dedicated electronic music VCO's. In particular, the VCO's from CES [1] and SSM [2] have made standard VCO design much simpler. Much of the theory of VCO design is available [3] and our most recent concentrated effort in VCO's goes back to our ENS-76 series from 1977 [4]. In particular, the ENS-76 Series VCO's Options 1 and 2 have been popular and highly praised by users. So why a new VCO design here? Well, there is still room for designs that offer features not available on single chip VCO's, and where an improved design can be offered. The design here uses a reversing VCO for through-zero FM capability. The design has been around in its basic form for several years, having been suggested to us by Douglas Kraul in EN#62 [5]. We have made some modifications, but in most cases, the important ideas were in Douglas' design. What we have tried to make is a VCO that offers through-zero modulation, and which is still well suited to standard VCO applications. Thus the present VCO improves on previous designs of through-zero devices [6,7,8] in the areas of accuracy and stability. It is also easy to build and has no expensive multiplier chips. It features as well a full set of standard waveshapers.

## THEORY OF OPERATION:

The through-zero VCO, (TZVCO) is based on the reversing oscillator principle. The block diagram of Fig. 1 will give the basic ideas. It is composed of a basic triangle/square oscillator based on the integrator-Schmitt-trigger principle. Here however the Schmitt trigger has the capability of switching state in a triggered manner, not just in response to the reaching of upper or lower trip levels. The Schmitt trigger will change output level when the input voltage reaches the trip levels, but it can also be triggered to reverse at will through the use of a triggerable flip-flop. All that remains to do then is to arrange to make the triggered reversals at the zero crossings of the linear control, and to rectify the linear control voltage.



Here we will not need to go into details about the FM synthesis process or the through-zero FM process. Information on FM synthesis can be found in [9] and a fairly recent summary of through-zero interpretations can be found in [10]. Consideration of these results leaves us only with the tasks of determining what a through-zero VCO should do, and how to implement these processes. We find that the waveform of an oscillator passing through zero should become level as zero is approached, and at zero control, it should then reverse direction (see again [10]). The system of Fig. 1 accomplishes these tasks. First, the reference current to the exponential stage is proportional to the magnitude of the linear control voltage. Secondly, at the time that the sign of the linear control changes, the Schmitt trigger is "flipped" to cause the oscillator to reverse direction.

The circuit diagram of the TZVCO is shown in Fig. 2, with Fig. 2a showing the oscillator circuitry, and Fig. 2b showing the added waveshaping circuitry. Most of the circuitry will likely be familiar to readers of this newsletter. Much of the circuitry not involving the through-zero process comes from ENS-76 VCO Option 2.

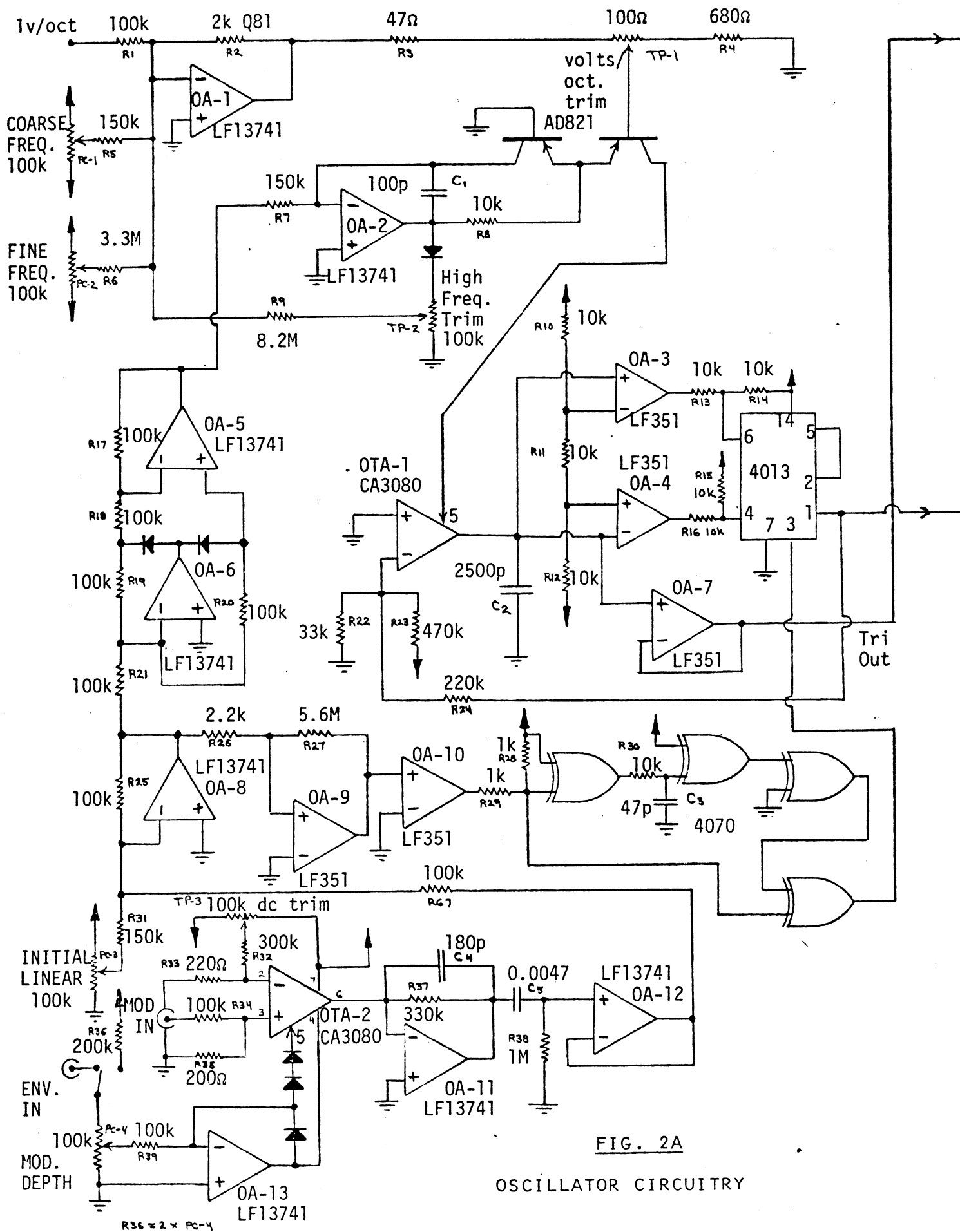


FIG. 2A  
OSCILLATOR CIRCUITRY



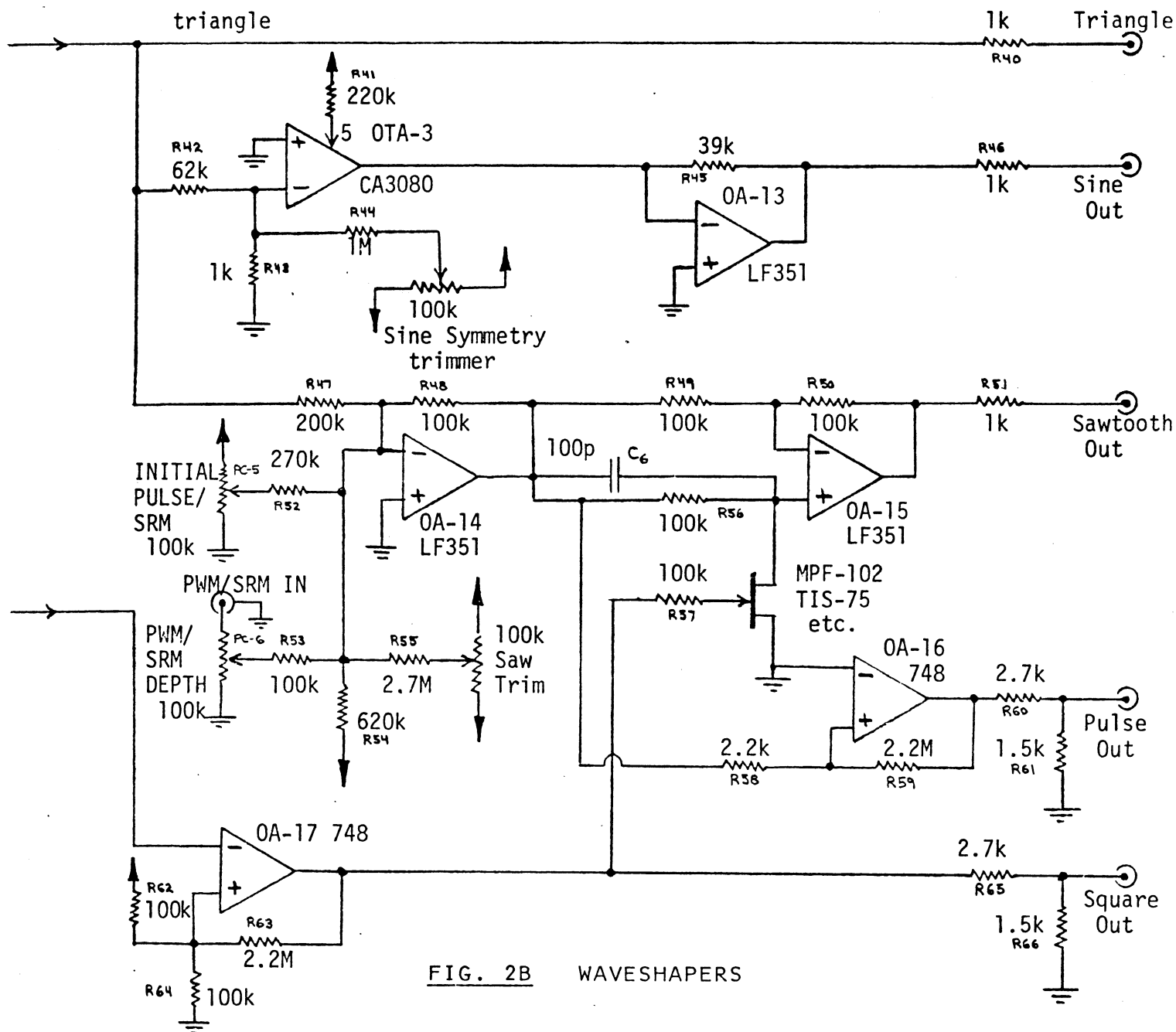


FIG. 2B WAVESHAPERS

The circuitry around OA-1, OA-2, and the AD821 matched PNP transistor pair, is our more or less standard exponential stage. The actual triangle oscillator is composed of OTA-1, OA-3, OA-4, OA-7, and the 4013 flip-flop. OA-6 and OA-5 form the full-wave rectifier for the magnitude of the linear control, as summed by OA-8. The sign of the linear control is determined by the op-amp Schmitt trigger (OA-9) and comparator (OA-10), with the circuitry around the 4070 quad exclusive-OR gate being the edge detector. The lower portion of Fig. 2a is an added section, not shown in Fig. 1. It is basically a built-in voltage-controlled amplifier for the purpose of handling a dynamic modulating signal. It is composed of OTA-2, OA-11, and OA-13, with OA-12 providing AC coupling for additional rejection of the control envelope. Fig. 2a above shows the remainder of the VCO circuitry, the waveshapers. OTA-3 and OA-13 form the sine shaper. OA-15 along with the FET below it is the heart of the triangle-to-saw converter. This configuration, as fed by the offset triangle from OA-14, provides the symmetrized ramp as well as the saw. The same driving waveform from OA-14 is fed to OA-16, providing a variable width pulse. OA-17 is an additional comparator forming the square wave from the flip-flop, with this square wave also being of the proper phase to control the triangle-to-saw converter above.

It will be useful at this point for the reader to study the circuit and sort out in his own mind the various sections, and to note in particular that the circuit is really two circuits in one. First, there is the more or less conventional VCO. Secondly, there is the additional circuitry for linear FM and through-zero FM. Roughly speaking, the upper portion of Fig. 2a (except OA-5 and OA-6) and all of Fig. 2b constitutes the conventional VCO. Thus the lower portion of Fig. 2a is the added linear FM circuitry. In fact, when either the modulating signal or the envelope is disconnected, the circuit behaves like a normal VCO.

Although it would be the usual practice to first describe the circuit, and then go on to applications, it will be useful here to show the basic sort of patch in which this circuit would be useful. This patch is shown in Fig. 3b, with Fig. 3a showing the block diagram of a through-zero FM system, with the blocks that are contained in the present TZVCO being within the dotted lines. It is possible to see that this TZVCO is a nearly complete through-zero system, lacking mainly the modulating VCO, but the depth-control VCA (VCA-2 in Fig. 3a) has been put inside. One could also include an internal VCO-2, etc., and this does not seem necessary. Avoiding the necessity of patching in VCA-2 each time however seems worth the trouble, so this is the purpose of the circuitry around OTA-2 in Fig. 2a. In addition, the dedicated internal VCA makes it possible for us to make certain additional control rejection features which are very important here. These will be described later.

## DETAILED DESCRIPTION OF

### THE CIRCUITRY:

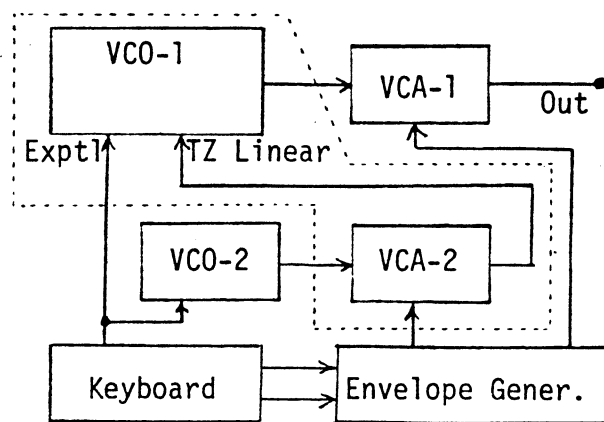


Fig. 3a Block TZ - FM System

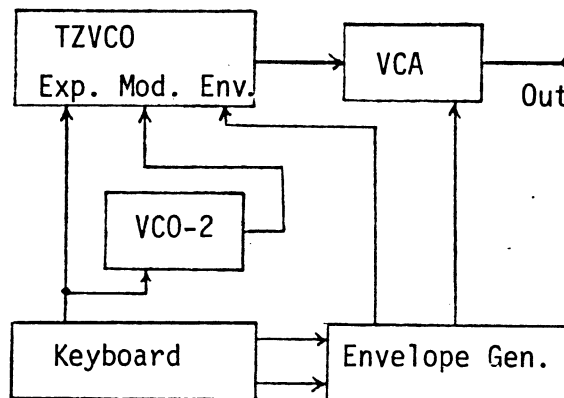


Fig. 3b Patch of the Present TZVCO

The basic exponential stage is shown in Fig. 2a around OA-1, OA-2, and the AD821 matched pair. We should point out a few features of this that have perhaps not been seen before, or which need review. We have not included an envelope input to this stage, but only the Coarse, Fine, and 1 volt/octave controls. If desired, the envelope input can be added exactly as in previous circuits [11]. The AD821 matched pair is a good choice, but unfortunately, Analog Devices is apparently not making this anymore. Many readers still have these in stock however, and we will be looking for a good replacement to suggest. Other good quality PNP pairs should work well. For a start, you can just use two unmatched transistors of the same type. In fact, I always build my circuits this way first and get them working, putting in the AD821 pair last after much of the danger of destroying it accidentally during construction and testing is over. The main problem with unmatched transistors will be thermal drift of frequency over time. Note that the 150k resistor attached to the (-) input of OA-2 is what would normally be the reference current set, running to -15 volts. Here it is attached to the full-wave rectifier, which provides a negative voltage proportional to the magnitude of the linear control sum. Finally, note the use of high-frequency compensation of the exponential stage through the diode attached to the output of OA-2 and through TP-2 and R9. This type of compensation, as suggested by Rossum [12], is similar to that of Moog and Hemsath [13]. Its theory and operation will be discussed later when we get to the tune-up procedures.

The actual oscillator circuit is composed of OTA-1, OA-3, OA-4, and the 4013 flip-flop, with OA-7 acting as a buffer for the triangle voltages. The oscillator is similar to others described [5, 14]. The OTA-1 serves as a current-controlled current source, driving the 2500 pf capacitor C2, thus forming an integrator. The voltage on the capacitor is fed to the buffer OA-7, and also, in parallel, to the two inputs to comparator op-amps OA-3 and OA-4. Although this means that there are three inputs connected to the capacitor voltage, the op-amps are low input bias types, and by using them in this parallel manner, we avoid any slowing down due to a buffer (that is, if OA-7 buffered the capacitor and then fed to OA-3 and OA-4). Note that the resistor string R10, R11, R12 sets the (-) input of OA-3 at +5 volts and the (+) input of OA-4 at -5 volts. The capacitor voltage normally will remain within these limits. If the capacitor voltage tries to exceed +5 volts, which would normally occur with the (-) input of the OTA-1 negative, then OA-3 will go high. This will set the flip-flop through pin 6. This will cause pin 1 of the flip-flop to go high (Q=1) and the (-) input of OTA-1 will be positive, causing the capacitor voltage to ramp negative. At the lower transition level (-5), a similar process occurs through a reset of the flip-flop through OA-4 and pin 4 of the flip-flop. This describes the basic oscillator which produces a triangle wave. Note for the moment that we do have pin 3, the clock pin of the flip-flop, that can also flip the output. This will be discussed later.

OA-8 is the linear control voltage summer. It sums an initial linear voltage from the pot with the output voltage from OA-12, which is the modulating signal. The full-wave rectifier around OA-6 and OA-5 does nothing more than give the negative of this linear control sum, thus modulating the reference current to the exponential control stage (through R7). We also need to know the sign of this linear control sum, and this is the purpose of OA-9 and OA-10. It was found experimentally that neither the single comparator (OA-10) nor the weak Schmitt trigger comparator (OA-9) would give reliable triggering over a wide frequency range. The combination works well however. Now, we are not interested in the direction of transition, but only in the fact that the sign changes. Thus we need an edge detector, and this is formed around a single package of exclusive-OR gates in the 4070 package. Basically the idea is the feed the two inputs of an exclusive-OR gate with the same signal, but to delay the path to one of the inputs. Thus the upper three EXOR gates along with the R30-C3 delay serve to delay the comparator output to the upper input of the lower EXOR gate. The EXOR has "one and only one" logic, and most of the time, the two inputs are the same, and the output is low. At the moment of transition, the direct path changes immediately, while the delayed path does not. Thus for a moment, there is one and only one, and the output goes high. After a moment (about a half microsecond), the two inputs become the same, and the output of the gate goes low again. This short pulse is the one that triggers the flip-flop, reversing the oscillator at the zero crossings of the linear control.

As mentioned briefly above, OTA-2, OA-11, OA-12, and OA-13 form the VCA to control the depth of the modulating signal. The circuitry is quite conventional [15, 16] and uses Jung's method [17] of driving the CA3080. In theory, there is no reason that we could not drive the (-) input of OA-8 directly from the current output of OTA-2. In practice however, it is advantageous to use OA-11 and OA-12, driving OA-8 through R67. The reason for this is that we have to be very careful with DC drifts and control envelope feedthrough, all of which can distract from the FM process. Our first step here is to zero the offset of OTA-2 using TP-3. OA-11 is a current-to-voltage converter. The 180pf capacitor is needed to remove some instability at a certain voltage level, which otherwise caused erratic triggering all the way through to the flip-flop. Next, we use a high-pass filter, C5-R38, to remove any residual DC feedthrough of the envelope. This filter, as buffered by OA-12, has a cutoff around 30 Hz, so modulating signals of a lower frequency are attenuated. [If desired, a switch to short out C5 for lower modulating frequencies can be used.] The overall circuit provides a clean DC-free controlled-depth modulating signal.

Fig. 2b shows the waveshaping circuitry as described briefly above. Since it is very similar to that presented in the ENS-76 Option 2 VCO [4], we need not take space here to review it. We will make a few comments on it when we discuss the tune-up procedures of the entire circuit.

## CONSTRUCTION AND TUNE-UP:

There should be little difficulty in the layout and construction of this circuit. The only unusual parts are the AD821 matched pair (discussed above) and the 2k Q81 resistor (+3500 ppm/°C), R2. The rest of the parts are fairly standard. The op-amps marked LF351 should be of that type or some equivalent BiFET of similar speed. The op-amps marked LF13741 are the type with BiFET input and type 741 output (slower). These are used where speed is not needed because they are cheaper in some cases. LF351's can be used for these if desired. In fact, LF351's can also be used for the two 748 op-amps if desired. The OTA's are type CA3080, and the two logic IC's (4013 and 4070) are standard CMOS IC's. Resistors are 1/4 watt 5% tolerance types, except for R1 which is better if made a 1% type. Panel pots and trim pots could actually have any value in the general range of 10k to 200k, and need not be the value shown as 100k (except for TP-1 which must be 100 ohms). If possible, C2 should be some sort of plastic film capacitor, and its value need not be exactly 2500pf (2000 pf to 3000 pf is good). The power supply lines should be bypass with 0.02mfd capacitors (not shown in the schematic) at two or three places in the circuit (for each supply). A good place for these is in the vicinity of the pulse handling circuits: OA-9, OA-10, OA-3, OA-4, and the 4013 and 4070 IC's. These capacitors are connected from the supply line to ground, and serve to keep spikes on the lines from moving to other areas of the circuit.

Naturally it is a good idea to (at least mentally) isolate and separately test the various parts of the circuit. When the circuit is turned on, it should begin oscillating immediately, producing a triangle wave at the output of OA-7. If it does not, connect a 33k resistor from pin 5 of OTA-1 to ground, first removing the line from the AD821 pair. If it still does not oscillate, there is a problem in the basic oscillator (OTA-1, OA-3, OA-4, 4013, OA-7). If it does work, then check out the exponential circuit at the top of Fig. 2a, and reconnect the line to pin 5 as in the original circuit. [Make sure there is some non-zero setting of the INITIAL LINEAR control (some negative voltage at the output of OA-5) as well, or the oscillator might be running very slowly.]

With the oscillator now running normally, we need to check the reversing circuitry, and there are several ways to do this. You should plan to make several of the tests to assure yourself that the circuit is working properly. For these tests, the INITIAL LINEAR control should be set about 1/4 of the way up. A signal (sine or triangle best) should be of  $\pm 5$  volt level and attached to MOD IN. Then the switch on the MOD DEPTH control should be put in the 200k resistor position, so that the control itself sets the depth (not depending on an envelope). The tests are as follows:

1. Set the oscillator to a comfortable frequency with the MOD DEPTH control all way down. Set the modulating frequency to some audible value (say 100 Hz to 200 Hz). Now, listening carefully to the output (sine or triangle of the VCO suggested), turn up the modulation depth slowly. You should hear the depth and complexity increase continuously with the control. There should be no points in the control setting where there is a sudden jump of output pitch or timbre, or other instability. Repeat the test for a variety of modulating and VCO frequencies.
2. Drop the VCO to a low frequency (1 Hz or lower) and the modulating oscillator even lower if possible. Turn the modulation depth all the way down. Observe the output waveform as it moves up and down between +5 and -5. Now turn the modulation depth all the way up. There should now be points in the output waveform where the direction changes within the limits of +5 and -5 rather than at them. (See Fig. 4a).
3. Turn the VCO and modulating frequencies back up, and the depth back down to zero. Listen to the square-wave output (or pin 1 of the flip-flop). The pitch should be the same as that of the triangle or course. Now, short out capacitor C2 with a jumper. The VCO should stop. Now turn up the modulation depth. At some point, an output will be heard from the square wave again. Check to see that this pitch is the same as that of the modulating frequency. Note the level

of the depth at which the square wave output starts up again. You may note that this is the same point at which the output of the full-wave rectifier changes from a sine wave to a rectified one (as in Fig. 4b). This is the position (for this setting of INITIAL LINEAR and of modulating signal level) where the modulation goes through zero. [Note that what is happening here is that the normal reversals have been prevented by the jumper on C2, and all that is getting through is the change-of sign pulses from the modulating signal.]

4. Leaving the INITIAL LINEAR control alone, you can now make a direct listening and visual test (oscilloscope waveform). Try to set the modulating frequency to something like 1/5 of the VCO's starting frequency. Observe the waveform and slowly turn up the depth. As you approach the through zero point (determined from 3 above), you will see "flat spots" appearing in the waveform (see Fig. 4c). As you go even higher, these flat stopping points will turn into actual reversals (see Fig. 4d). You should be able to identify these reversal points with the zero crossings of the modulating waveform. Once this all checks out, again do a careful listening test passing through the zero point. Note that nothing drastic happens. The through-zero case is just a continuation of the process that was going on above zero. If you do hear a drastic change, then something is wrong. Probably the flip-flop is not triggering on both edges.

With the oscillator working properly in both its normal mode and its through-zero mode, it is time to check out the envelope circuitry. To do this, connect up a voltage of about +5 volts to the MOD IN (or use 200k in series with the input and connect to +15). Trigger an envelope generator from a low-frequency oscillator for a periodic envelope, and connect this envelope to ENV IN. [You can also trigger envelopes by hand by pressing a key on the keyboard if this is convenient.] Now, short out capacitor C5 with a jumper. Listening to the VCO output, adjust TP-3 for minimum change of pitch during the envelope cycle. [Or you can observe the DC feedthrough directly at the output of OA-11]. Finally, remove the short from C5, and the feedthrough of control envelope should be minimal.

The waveshapers should be simple to adjust. The sine shaper is adjusted with the symmetry trimmer for the same shape top and bottom. There may be a slight suggestion of the triangle's point at the top, but this is normal for the lowest distortion level with this shaper [18]. This oscillator has the Pulse-Width Modulation (PWM) and Symmetrized Ramp Modulation [SRM] features previously reported [4, 19]. With the INITIAL PULSE/SRM control all the way down or all the way up, the waveform at the saw output should be very close to sawtooth. With the control all the way down, adjust the Saw Trim control for the best possible alignment of the sawtooth segments. [If this does not work out, a likely problem is that the FET is connected wrong. You can try all possible combinations of the FET's leads without any danger to the FET or the circuit.] The pulse output should respond to the initial pulse control. Note that the pulse width will be zero or close to it with the INITIAL PULSE/SRM control at zero or at maximum, so assure that it is somewhere in the middle when testing the pulse output. The square waveshaper needs no adjustment and is obviously working if the sawtooth is. If it is not working somehow, then the sawtooth converter won't work either.

The circuit should now be working, and it is time for the final tune up of the exponential response. This is done by adjusting TP-1 and TP-2. To set up for this, you will need a frequency counter or a good ear, or perhaps you can rig up a flip-flop (e.g. the unused side of the 4013) to provide an octave reference. Set a reference voltage to -1.000 volts. Set the coarse and fine frequency controls to their center zero position, and set the INITIAL LINEAR control so that the output of OA-5 is about -10 volts. Disconnect any modulation signals. The frequency of the VCO should be about 1200-1300 Hz. Connect the -1.000 volt signal to the 100k input (R1) and the frequency should drop to about half. Assure that TP-2 is in its lower position and adjust TP-1 until the -1.000 signal gives you exactly half the original frequency. Adjust the input voltage to +4.00 (4.000 if your meter will do this). Multiply the zero volt frequency by 16. The actual frequency will likely be a bit on the flat side of this target value. If so, adjust TP-2 upward until the correct value is reached.



Fig. 4a Low-Freq. Reversals

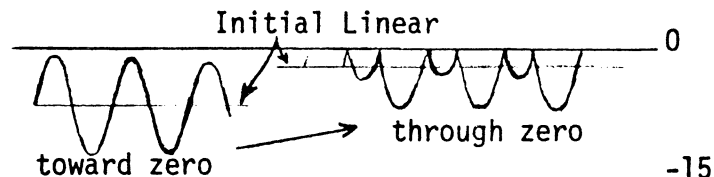


Fig. 4b Full-Wave Rectifier Output



Fig. 4c Near-Zero Approach



Fig. 4d Through-Zero

Now it is up to you how much more work you want to do on this calibration. If you have the desire, equipment, and time, you can run the control voltage up and down, and make minor adjustments of TP-1 and TP-2 until you get a very accurate response function. The values below for my oscillator will give you some idea of what can be achieved with about a half-hour's playing around.

| <u>VOLTAGE</u> | <u>WITH TP-2 AT BOTTOM</u> |              | <u>WITH TP-2 ADJUSTED UPWARD</u> |              |
|----------------|----------------------------|--------------|----------------------------------|--------------|
|                | <u>FREQUENCY</u>           | <u>RATIO</u> | <u>FREQUENCY</u>                 | <u>RATIO</u> |
| -5.00          | 39                         | 2.000        | 39                               | 2.000        |
| -4.00          | 78                         |              | 78                               |              |
| -3.00          | 156                        | 1.994        | 156                              | 2.000        |
| -2.00          | 311                        |              | 312                              |              |
| -1.000         | 622                        | 2.002        | 625                              | 1.992        |
| 0              | 1245                       |              | 1248                             |              |
| +1.000         | 2493                       | 1.989        | 2499                             | 2.000        |
| +2.00          | 4959                       |              | 4998                             |              |
| +3.00          | 9844                       | 1.998        | 9994                             | 1.997        |
| +4.00          | 19668                      |              | 19959                            |              |

The above values do indicate that the circuit is quite satisfactory over the normal audio range. The actual range of the oscillator is from well below 1 Hz to about 29,000 Hz. Note that the curve for the uncompensated oscillator first goes flat, and then starts to go sharp again, a strange behavior seen in a similar oscillator [4]. Since we can only measure frequency to the nearest Hz, you should note that the value of frequency for the compensated case, -1.000 volts, might have been 624 Hz instead of 625, and this would have made the response nearly perfect. Probably the most important feature of the TP-2 high frequency trim is its improvement in the 5000 - 10000 Hz octave. It should be noted that some drift at the 20,000 Hz level was noted. The frequencies were initially a bit higher than those listed above, but drifted downward to those listed after a minute or two. Use of such high frequencies is rare anyway, but you might make a mental note that for occasional short high notes, the results may be a bit sharp. The best guess is that some sort of self-heating is going on at these higher frequencies (higher controlling currents). It is not a real problem.

#### APPLICATIONS:

If you have never heard dynamic-depth or through-zero FM tones before, you will be pleased with the wide range of interesting sounds you will get from this simple circuit. Note however that through-zero is an extension of dynamic depth. In conventional linear modulation, we are able to use dynamic depth, but the instantaneous frequency can only approach zero, so the peak depth is limited. A through-zero device can have a much larger peak depth, since it does not have to stop at zero. But it must be realized that going through zero in itself is not dramatic. The difference between going to within 1% of zero, and going through zero by the same 1%, is not much different than that between 1% and 2% on the positive side.

The principal application of this oscillator will be to the production of tones with dynamically varying spectra. If the ratio between the original VCO frequency (called the carrier) and the modulating frequency is an exact integer, then the resulting through-zero FM tone will be harmonic. Otherwise (the general case), it will be composed of non-harmonic partials. The latter case is very useful for the production of bell-like and other percussive sounds. The basic patch here is that of Fig. 3b. Note that you can easily set this up as the basic voltage-controlled patch, leaving out the VCF. [Of course you can use the VCF if you want, but for now you will just want to hear the FM process itself.] Once the patch is working in its basic form, you just plug a VCO output into the MOD IN, and an envelope (the same, or better, a different one than that going to the VCA) into ENV IN. You are now ready to explore the different timbres that can be achieved. You will want to spend some time with this patch.

It should be noted (see Fig. 5) that in general all modulated tones will begin with modulation above zero, and go through zero as the depth increases, with the modulating voltage increasing about an Initial Linear level with the depth. Thus by selecting the depth and the Initial Linear, we can modulate over any range we please. However, note that the original frequency (carrier) of the VCO changes with the Initial Linear setting (since that is the VCO's reference current). What this means is, for example, that if we want the VCO's frequency to go further through zero, we can increase the depth (obviously) or we can lower the Initial Linear. If we increase the depth, then the total range swept is increased. If we lower the Initial Linear, the original pitch drops. This latter case is no real problem, as we can simply raise the exponential control back to give the original pitch. Therefore, there are no real restrictions here. However, as a matter of convenience, it might be nice if we could change the Initial Linear and have the VCO carrier remain the same. A way of

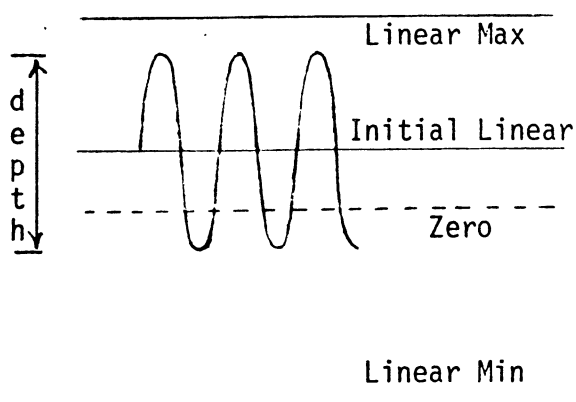


Fig. 5 Effect of Initial Linear and Mod Depth controls

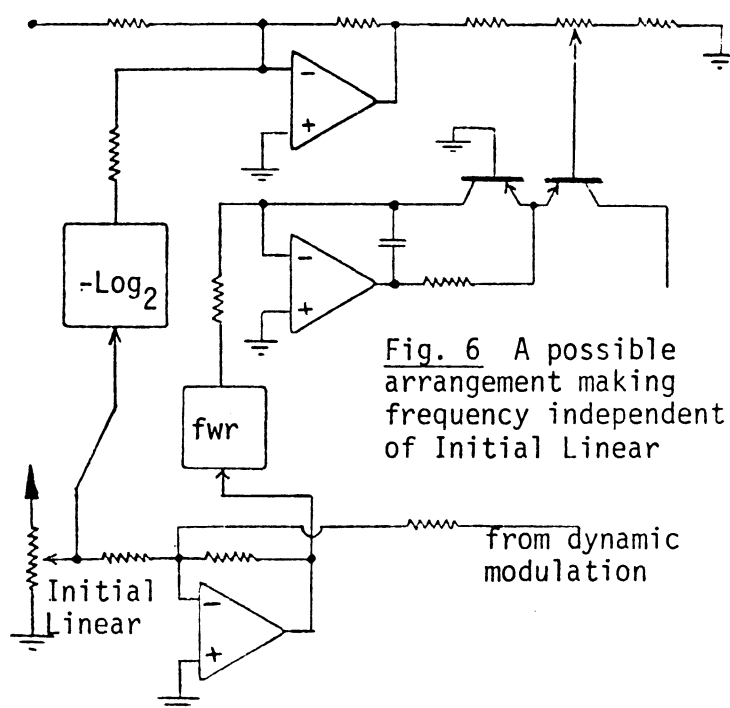


Fig. 6 A possible arrangement making frequency independent of Initial Linear

achieving this is shown in Fig. 6. We just need to add a log converter. Thus for example, if the Initial Linear were changed from 8 volts to 4 volts, the carrier would normally change to 1/2 its original frequency. But with the added  $\log_2$  loop, we would get a change of +1 volt back at the exponential converter, moving the carrier frequency back up one octave. There is probably some neat way of doing this. In particular, it looks as though we could avoid the path through the upper op-amp of Fig. 6 and just connect the base of the left transistor of the exponential converter (here shown grounded) to a base of one of the transistors in the log converter. The details of this and other possible developments may be reported in a later issue.

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## MANUALLY-CONTROLLED OSCILLATORS

MANUAL OSCILLATORS: Manually controlled oscillators, often used as low-frequency control oscillators (LFO's), are often used in synthesizer systems. In this collection, we will first give three application notes which describe the construction of some basic oscillator types: sine, triangle-square, and sawtooth. In addition, a recently reported LFO circuit by Pete Lutz is reprinted, along with a circuit using the 8030 chip from Craig Anderton. Finally, we include a Sine-Square oscillator of Ralph Burhans, and a circuit in response to a reader question.

| Preferred<br>Circuit<br>Option | Original<br>Source | Original<br>Designer | Estimated<br>Parts<br>Cost | Notes                         |
|--------------------------------|--------------------|----------------------|----------------------------|-------------------------------|
| MCO-1                          | AN-29              | Hutchins             | \$2-\$3                    | Sine Oscillators              |
| MCO-2                          | AN-67              | Hutchins             | \$2                        | Triangle-Square               |
| MCO-3                          | AN-79              | Hutchins             | \$2-\$3                    | Sawtooth                      |
| MCO-4                          | EN#88              | Lutz                 | \$5                        | Ramp/tri/square/<br>pulse out |
| MCO-5                          | EN#53              | Anderton             | \$9                        | Squ/tri/sine                  |
| MCO-6                          | EN#44              | Burhans              | \$3                        | Sine/Square                   |
| MCO-7                          | EN#83              | Hutchins             | \$2                        | Triangle-Square               |



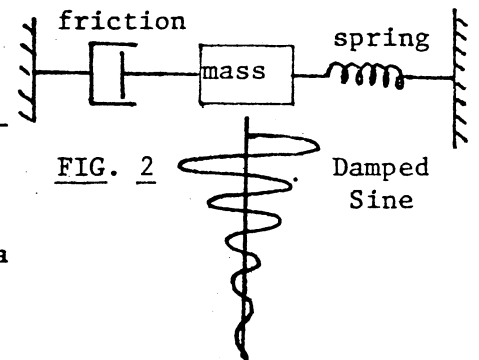
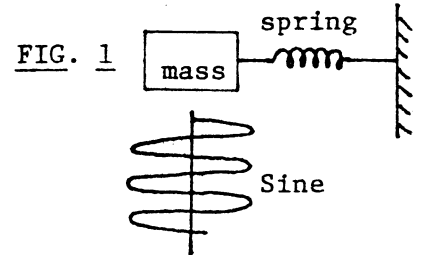
ELECTRONOTES MUS. ENG. GROUP  
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 (607)-273-8030

APPLICATION NOTE NO. 29

March 5, 1977

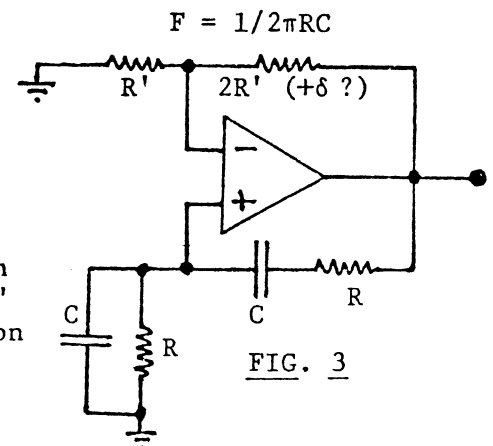
SIMPLE SINE WAVE OSCILLATORS

It should be easy to make a sine wave oscillator, at least it seems that way. After all, the sinusoidal response is the natural response of many many physical systems. To see why it is not really so easy, let's approach the subject by considering how we would make a mechanical sine wave oscillator. Simple! We just use the setup of Fig. 1 where we use a spring and a mass and a frictionless surface. We give the mass either an initial displacement or an initial velocity and it gives us a sine wave from then on. You order your parts, but find that your frictionless surfaces have been back-ordered (indefinitely)! So you have to make due with your surfaces with friction (Fig. 2). Unfortunately, this system damps out. So now we see what we have to do. We have to put in some energy to make up for that which is lost. We just have to give the mass a little extra push once in a while - well let's be a little more careful about that. If we insert energy (or displacement) in any way except a way that would replace energy exactly as it is lost, we put little nicks and jogs in the waveform and introduce distortion. But, we might suppose that the friction may be very small so that there is only a very slight damping, so that we need only insert a very small amount of energy, perhaps once each cycle, and this will only produce a minimum of distortion. Sounds good, but the problem is that if a little too much energy is put in each cycle, the amplitude builds up, and if it is a little low, the amplitude will die away, even though at a much slower rate. Thus, we need some sort of amplitude regulator. It is exactly like pushing a child on a swing. If you don't push enough on each swing, the child will complain that he is not getting a good enough ride. If you push too much, you will become alarmed that the child will fall off. Thus you regulate for a normal swinging amplitude.



Before going on to actual sine wave generators of the regulated type we suggest above, we should note that it is always possible to generate something simpler like a triangle, and shape it to a sine-like form. This needs no regulation (assuming common type triangle oscillators) and is often used. Below we are looking for simple, mainly fixed frequency sine wave oscillators that will generally result in somewhat lower distortion than is possible using a triangle-to-sine waveshaper.

Probably the simplest sine wave oscillator is the so-called Wien Bridge Oscillator (WBO). The most basic form of this oscillator is shown in Fig. 3. It can be shown that the series and parallel R-C combination feeds back a voltage of  $1/3$  the output at a frequency  $1/2\pi RC$ . Thus we set the op-amp for a gain of 3, and we have (ideally) the perfect oscillator. If you build this as shown, when you first turn on the power supply, nothing will happen. What will probably happen is that oscillation will build up very slowly until the output "bumps its head" on the power supply rails. This will cause some distortion at the peak. If nothing seems to be happening, add a resistor in series with  $2R'$  to increase its value 1-2%.

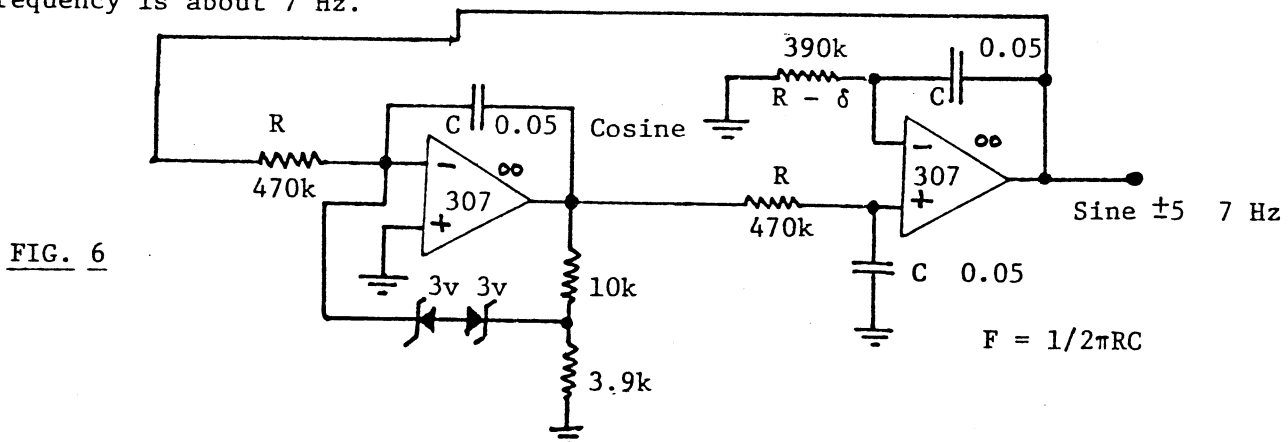


The WBO as shown in Fig. 3 may be fine in some applications, but it has several drawbacks: The distortion may be too high, it only gives an output amplitude limited by the supply, and component drift may cause it to stop oscillating, or to produce more distortion. For more exacting results, we will increase the gain of the op-amp slightly and add soft limiting with zener diodes as shown in Fig. 4. This arrangement gives an output voltage of  $\pm 10$ , although this may need to be trimmed up a little by changing either the 10k or the 9.1k resistor a little. If you want more amplitude, decrease the value of the 9.1k resistor, and if you want less, increase the value of the 9.1k resistor. If you need to make the oscillator variable, you can use a dual pot to vary the two R resistors, and the amplitude can be varied using a pot arrangement as shown in Fig. 5. With this arrangement, the frequency can be varied over a range from 10:1 to 100:1, depending on the matching of the pot sections, and the amplitude can be varied from about about a one volt level up to a 15 volt level. This is really not good enough for a "Function Generator" type of oscillator, but may be good enough for some simple experiments requiring a relatively good sine wave over a limited amplitude and frequency range. Whether fixed or variable, it is best to use well matched capacitors (1% or better) and the resistors should also be matched as well as possible. The op-amp chosen should have sufficient slew rate for the amplitude and frequency required. The maximum slew rate required is:

$$SR = 2\pi F \cdot A$$

Where F is the frequency (or max. frequency) and A is the amplitude (or max. amplitude).

Along with a wealth of other information on sine wave oscillators, Walter Jung describes a simple quadrature (Sine and Cosine outputs - 90° out of phase) oscillator in "The Signal Path: II - Sine Wave Oscillators," in *db*, July 1976, page. 38. Below we make minor changes in the circuit so that the amplitude output is  $\pm 5$  volts, and so that the frequency is about 7 Hz.



The oscillator is basically a single integrator and a double integrator in a loop. Note that one of the R resistors is made slightly smaller to assure that the loop starts up. The loop gain regulation is similar to that in the WBO above. We have in mind two applications for this circuit. First, 7 Hz is just about right for a musical vibrato generator, and it might be interesting to examine vibrato of two processes where the different signals are out of phase by 90°. Secondly, a quadrature signal of about 7 Hz is needed in many frequency shifting systems. In these schemes, signals to public address amplifiers are first frequency shifted slightly to prevent positive feedback through the air.

## ELECTRONOTES

APPLICATION NOTE NO. 67

1 PHEASANT LANE

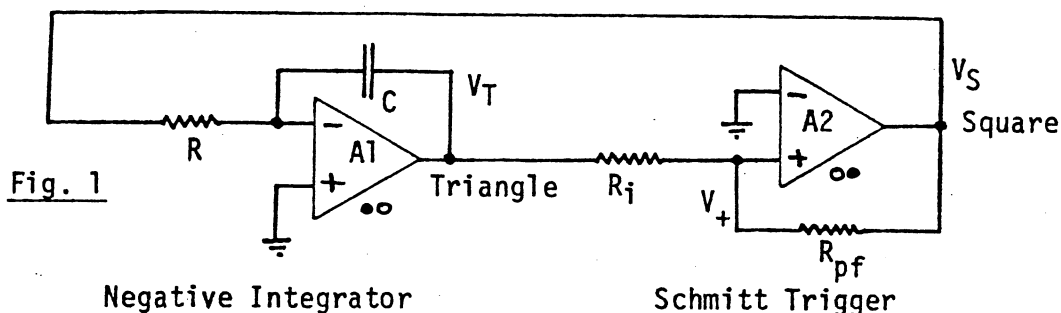
December 18, 1977

ITHACA, NY 14850

SIMPLE TRIANGLE-SQUARE OSCILLATOR

The simple op-amp multivibrator discussed in AN-28 uses only one op-amp and is relatively easy to use if only a square-wave output is needed. The curved triangle-like waveform that also appears in the circuit is completely unbuffered, and is therefore difficult to use in many applications. The present circuit produces a well-defined triangle waveform. It uses two op-amps and is essentially a negative integrator and a Schmitt trigger in a closed loop. The basic circuit is well known. Here we will review it, set up the basic equations, and show how some refinements in the calculations can be used to allow for the use of real op-amps.

The basic Triangle-Square oscillator is shown in Fig. 1.



The analysis of Fig. 1 starts with two assumptions. First we assume that since there is no negative feedback on A2 that the output of A2 is either at +15 or -15 (assuming a  $\pm 15$  volt supply), and secondly that since there is negative feedback on A1, that the (-) input is at ground potential along with the (+) input. These two assumptions tell us that there is a current through R of magnitude  $15/R$ , which may be either toward or away from the capacitor C. Since the (-) input of A1 is a very high impedance (ideally, it draws no current at all), the current through R must be flowing through the capacitor C (where else would it go?). Thus, the capacitor C is being charged and correspondingly the voltage across it is changing. Since negative feedback is being employed, the (-) input of A1 remains at ground potential and therefore the output voltage of A1 must be changing. We can consider the manner in which a capacitor charges. The voltage on a capacitor is related to the charge (Q) on it by  $Q = CV$ . Since C is a constant, the voltage and the charge are proportional, and hence the change of charge and the change of voltage are proportional as well. Now, the change of charge is the current into or out of the capacitor. We can represent small changes by the symbol  $\Delta$ . Thus  $\Delta Q$  is a small change in charge and  $\Delta V$  is a small change in voltage. We can assume all these changes to take place over a small interval of time  $\Delta t$ . The rates of change of charge and voltage are thus  $\Delta Q/\Delta t$  and  $\Delta V/\Delta t$  respectively. Since  $Q = CV$ , it is also true that:

$$\text{Current Through R} = \frac{\Delta Q}{\Delta t} = C \frac{\Delta V}{\Delta t} = C \cdot [\text{Slewing Rate of Triangle}]$$

Thus, for a constant magnitude of current, the output voltage of A1 changes at a constant rate.

We have assumed that A2 is either at +15 or at -15 volts. In either case, we are interested in the conditions under which the output of A2 will change. The circuitry around A2 is that of a Schmitt Trigger as discussed in AN-31. We can think of A2 as operating as a comparator, and since the (-) input is grounded, the output will change

whenever, the voltage on the (+) input crosses zero. We can consider  $R_i$  and  $R_{pf}$  to form a voltage divider. We assume no current is drawn by the (+) input of A2. Thus the voltage on the (+) input is:

$$V_+ = \frac{V_T R_{pf} + V_S R_i}{R_{pf} + R_i}$$

which is equal to zero when  $V_T R_{pf} = -V_S R_i = -15 R_i$ . This value of  $V_T$  is such that the (+) input of A2 is zero, and corresponds to the peak value of the triangle since beyond this point the output of A2 will switch, and the direction of change of  $V_T$  will reverse. Thus, we can get an expression for the amplitude of the triangle as:

$$V_{TM} = 15 \frac{R_i}{R_{pf}} \quad [\text{Maximum value (amplitude) of } V_T]$$

For one full cycle of the triangle, the voltage at the output of A1 must change through a full range of  $4V_{TM}$  (up, back to zero, down, back up to zero). We are now in a position to calculate the time for one cycle, and thus the corresponding frequency. We previously found the slewing rate of the triangle to be:

$$\Delta V / \Delta t = (\text{Current Through } R) / C = (15/R) / C = 15/RC$$

To use the old terminology for "Hertz" as "cycles-per-second", we can write:

$$\frac{\text{Cycles}}{\text{Second}} = \frac{\text{Volts}}{\text{Second}} \cdot \frac{[\text{Volts}]^{-1}}{[\text{Cycle}]}$$

Volts/Second is just  $\Delta V / \Delta t$  and Volts/Cycle is just  $4V_{TM} = 60R_i / R_{pf}$ , so we can arrive at the equation for the frequency of the oscillator of Fig. 1 as:

$$f = (15/RC) \cdot (60R_i / R_{pf})^{-1} = \frac{1}{4RC} \frac{R_{pf}}{R_i} \quad [\text{Frequency of Oscillation}]$$

Note that for oscillation to occur,  $R_{pf}$  must be greater than  $R_i$ , so the ratio  $R_{pf}/R_i$  must always be greater than 1. For normal safe amplitude levels, the ratio will be at least 2 or more. A 5 volt amplitude of the triangle gives  $R_{pf}/R_i = 3$ . Thus, the frequency of oscillation in most cases will be of general magnitude  $1/RC$ . Note that if  $R_{pf}/R_i$  is set exactly to 4, the frequency is  $1/RC$  and the amplitude is 3.75 volts.

Fortunately, for a moderate range of frequencies, 1 Hz to 2 kHz say, the above analysis works quite well. About the only change we need make is an adjustment for the fact that real op-amps may not reach the full supply voltage, and thus the output of A2 will not be at a 15 volt level, but often more like 14 volts. This changes the amplitude formula to  $V_{TM} = 14R_i / R_{pf}$ , but the output voltage of A2 cancels out of the frequency formula so that stays the same. For normal op-amps,  $R$  should not be more than 1 megohm, and  $R_i$  and  $R_{pf}$  should be chosen in the range of 10k to 100k.

When it comes to very low frequencies (below 1 Hz), it is sometimes necessary to use an op-amp for A1 that has very low bias current, and perhaps to correct for input offsets of A1 and A2 according to the data with the op-amps. Note that in general you can not just make  $C$  larger to get lower frequencies because eventually you will need to use electrolytic capacitors which will have polarity and leakage problems. There are plenty of good and inexpensive op-amps available for A1 however (the CA3140 for example).  $R$  can be 100 megohms or more with such op-amps.

At high frequencies (greater than 2 kHz) the speed of A2 is likely to become a problem. Here, an uncompensated LM301 or type 748 is suggested for frequencies to 10 kHz. Above that, a high speed comparator should be used according to the setup data specified by the manufacturer. The 301 and 748 type op-amps may be used up to 30 kHz if waveform inaccuracies and inaccuracy in the frequency formula can be tolerated.

1 PHEASANT LANE

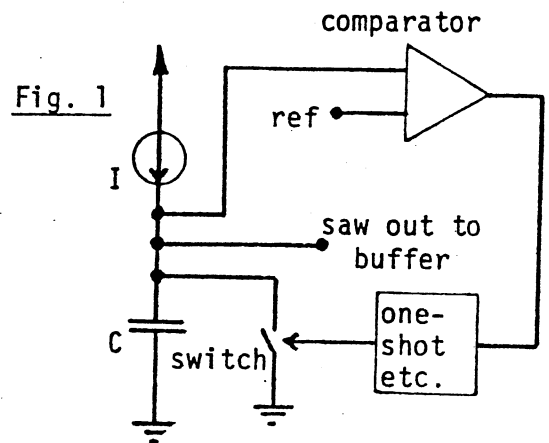
March 18, 1978

ITHACA, NY 14850

SOME SIMPLE SAWTOOTH WAVE GENERATORS

(607)-273-8030

The usual type of sawtooth wave generator consists of a source of constant (or nearly constant) current, a capacitor which is charged by this current, a device for detecting an upper level for the capacitor voltage, and a means of resetting the sawtooth by a rapid discharge of the capacitor. A typical setup is indicated in Fig. 1. It is important to realize that the reset portion of the scheme is generally a two-step device. One device is just the actual switch (transistor, FET, etc.) while the other part makes sure the switch remains closed long enough to sufficiently discharge the capacitor.



A typical sawtooth cycle would be as shown in Fig. 2a. It is interesting to consider what happens when the one-shot, or other timesetting device is left out. When the comparator fires on in such a case, the switch closes and the capacitor starts to discharge. However, very soon its voltage drops just below the reference level, and the comparator fires off, opening the switch. The resulting waveform is thus as shown in Fig. 2b. Many designers have discovered this by accident! Without the one-shot, the only delay is due to the speed of the comparator, and that is often fast enough that complete discharge simply does not occur.

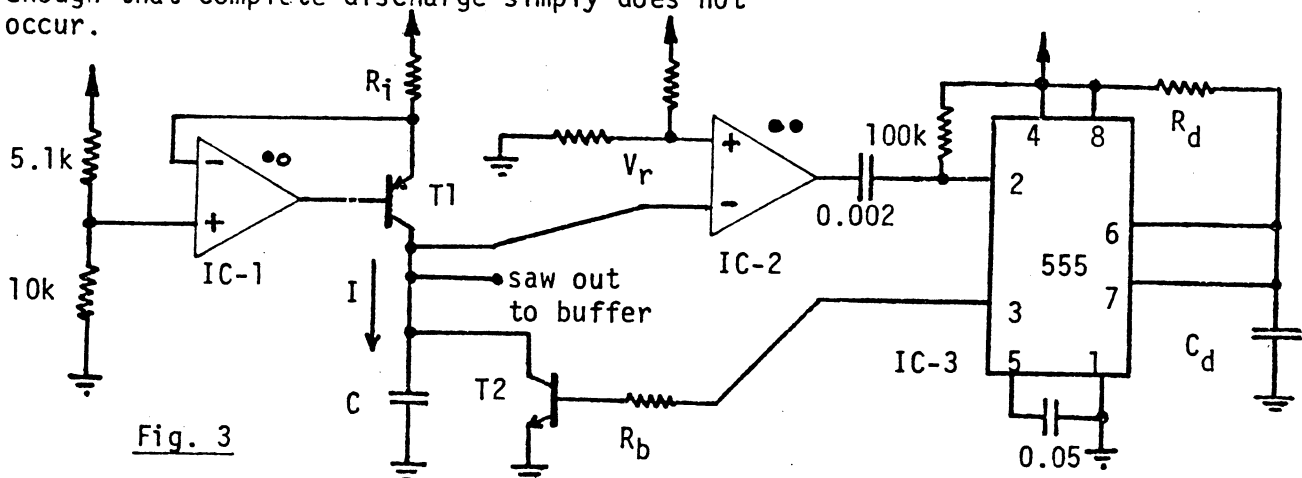
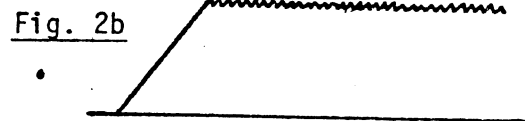
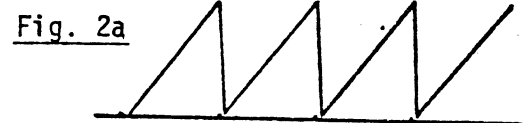


Fig. 3 shows a realization of Fig. 1 with actual components. This circuit has a few more parts than some other circuits, but it is very easy to control the different design parameters. The (+) input of IC-1 is at +10 volts, as must be the emitter of T1, and hence  $I = 5/R_i$  and since  $I = dQ/dt = CdV/dt$ , the rate of change of voltage on C is  $I/C = 5/R_i C$ . The reference voltage  $V_r$  on the (+) input of IC-2 is set by the voltage divider, and determines the sawtooth peak. When the voltage on C exceeds  $V_r$

the comparator IC-2 goes low, triggering pin 3 of IC-3 high for a time determined by  $1.1 \cdot R_d C_d$ , and +15 volts is applied to the base of T2 through  $R_b$ , causing the capacitor to discharge for that period of time. In general,  $R_d$  and  $R_b$  would be adjusted so that the sawtooth gets down below 1% of its peak value upon discharge, but does not remain down longer than is necessary.

In many cases, a somewhat simpler circuit will do the job, and a good example of such a simple circuit is shown in Fig. 4. This circuit is a simple variation on the standard triangle wave generator (see AN-67). Here, a diode and resistor are added.

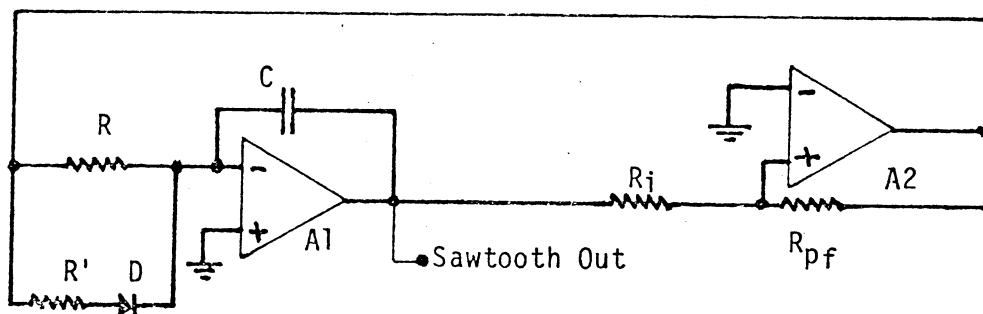


Fig. 4

Consider that when the output voltage of A2 is at its negative value (approaching -15), the diode is back biased, so no current passes through it, and integrator A1 ramps upward as current passes through resistor R. However, when the ramp reaches a certain upper value ( $15R_i/R_{pf}$ ), the output of Schmitt trigger A2 goes positive, and current can now flow through both R and through the series combination  $R'$  and diode D. If  $R'$  is much less than R, the integrator will ramp down very rapidly to the lower threshold of the Schmitt trigger, at which point, the output of A2 will again go negative, and the upward ramp will begin again. The value of  $R'$  should be in the range of 300 - 1000 ohms, and can be determined by experiment. It is not a good idea to make  $R'$  too small, or to leave it out, as strange behavior may occur that will depend on the individual diode and on the op-amps used. Note that if the diode is reversed, we simply obtain a falling ramp sawtooth instead of a rising one. The frequency may be obtained to a good approximation by doubling the frequency of the corresponding triangle wave as determined in AN-67. The result is:

$$f = \frac{1}{2RC} \frac{R_{pf}}{R_i}$$

When a high performance sawtooth generator is needed, the circuit of Fig. 5 can be used. Like the circuit of Fig. 4, this one uses a comparator with hysteresis to provide the necessary discharge duration. In Fig. 5, the necessary hysteresis is provided by the small capacitor in the positive feedback loop. The actual discharge

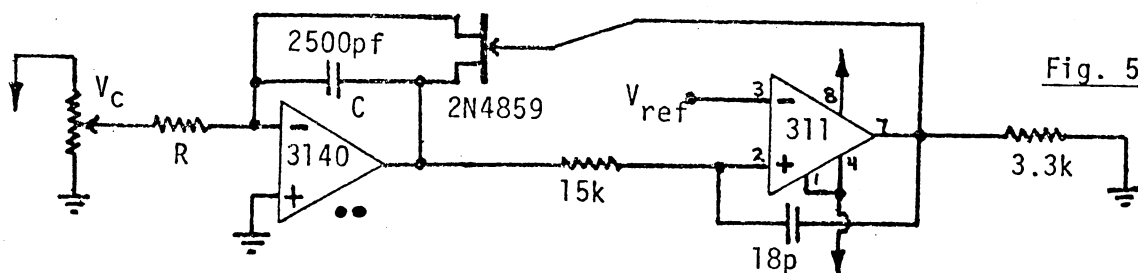
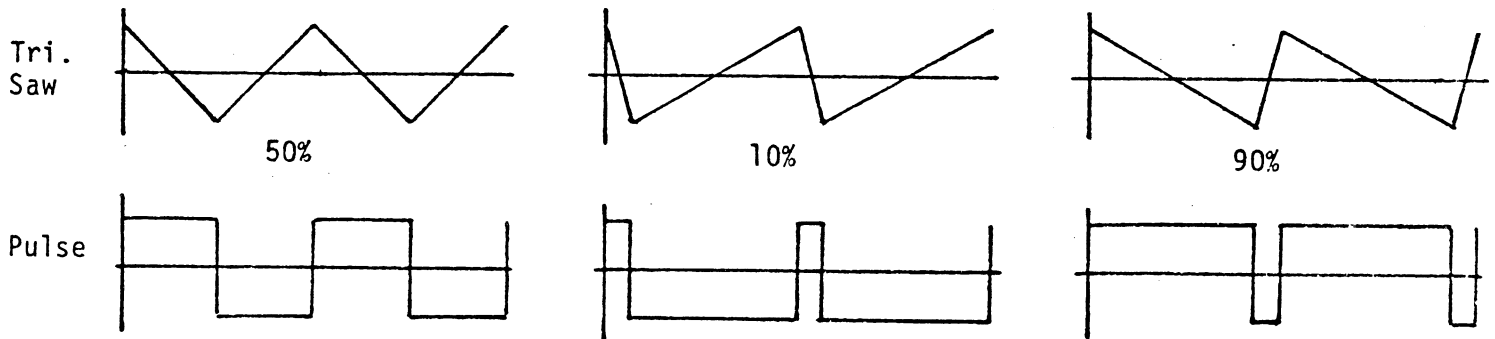
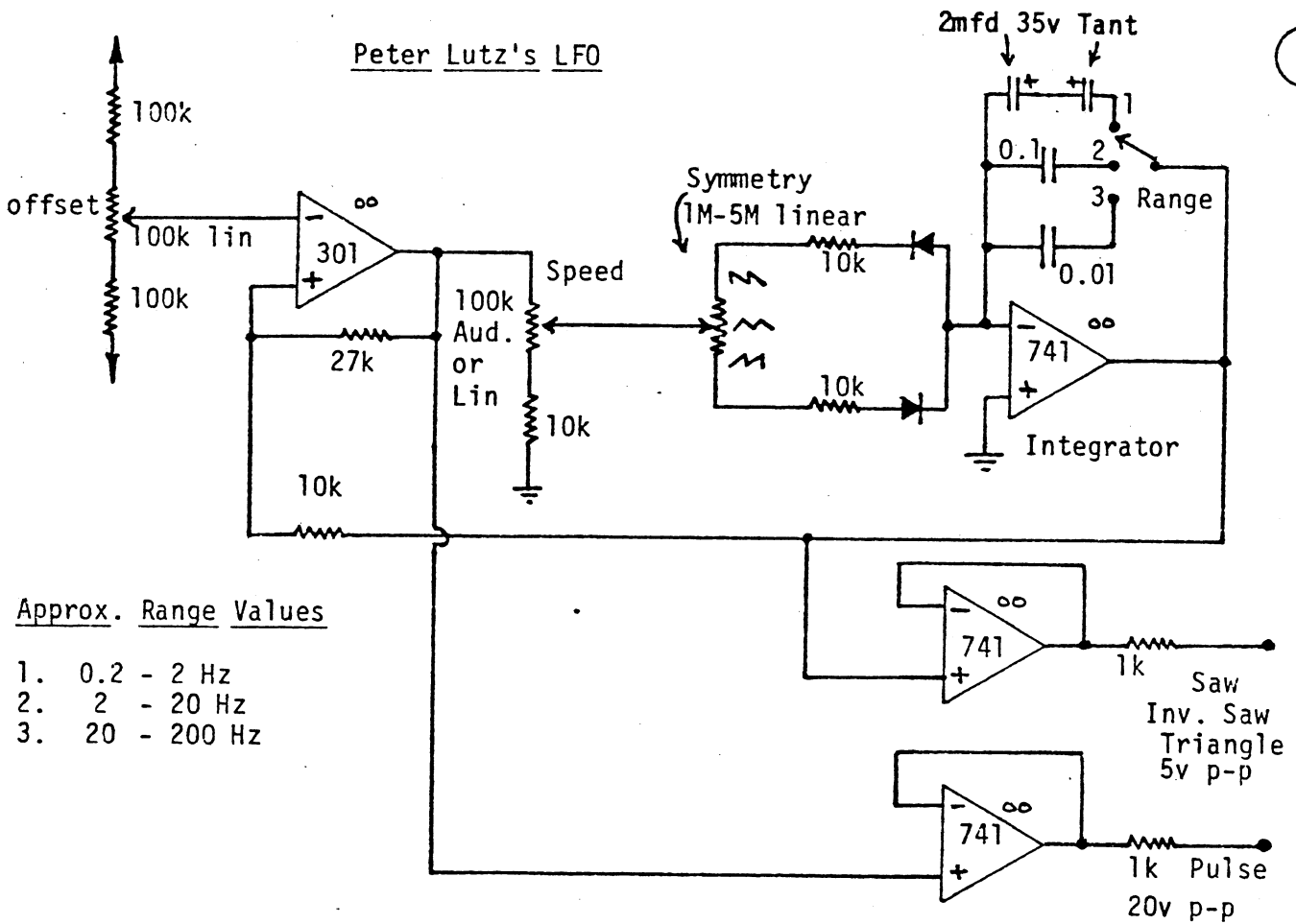


Fig. 5

is done by the 2N4859 switching FET, which discharges the 2500pf capacitor rapidly and cleanly. The input current through R determines the frequency. R may be replaced with any desired current sink.  $V_{ref}$  may be set with a voltage divider on the (-) input of the 311 comparator, and determines the peak amplitude. Be sure to note the unusual pinout of the 311 as compared to standard op-amps. The circuit of Fig. 5 is excellent as the heart of a voltage-controlled oscillator since the frequency is determined by the current into the (-) node of the CA3140 which forms the integrator. A wide range of frequency is available from this circuit designed by T. Mikulic.

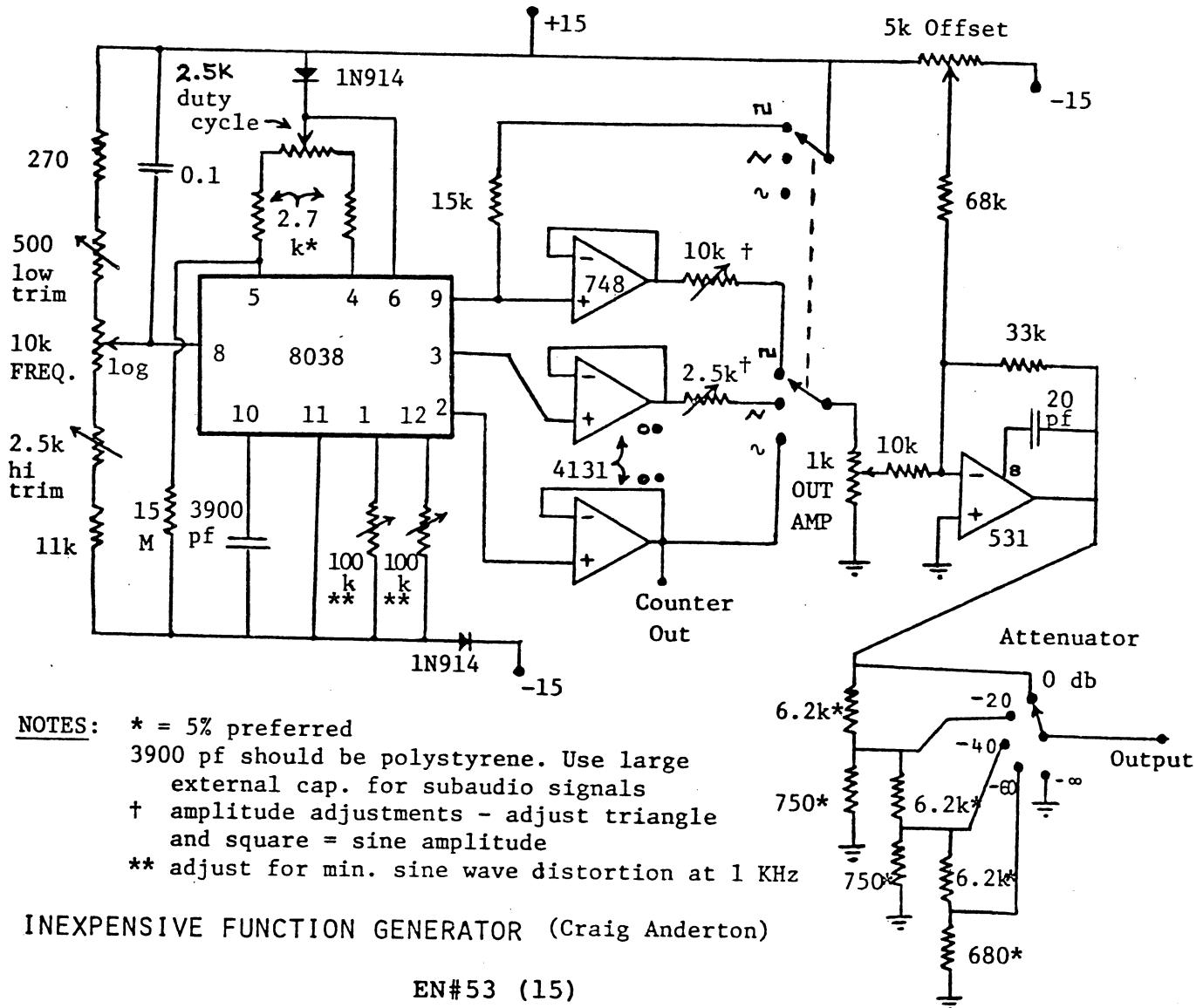


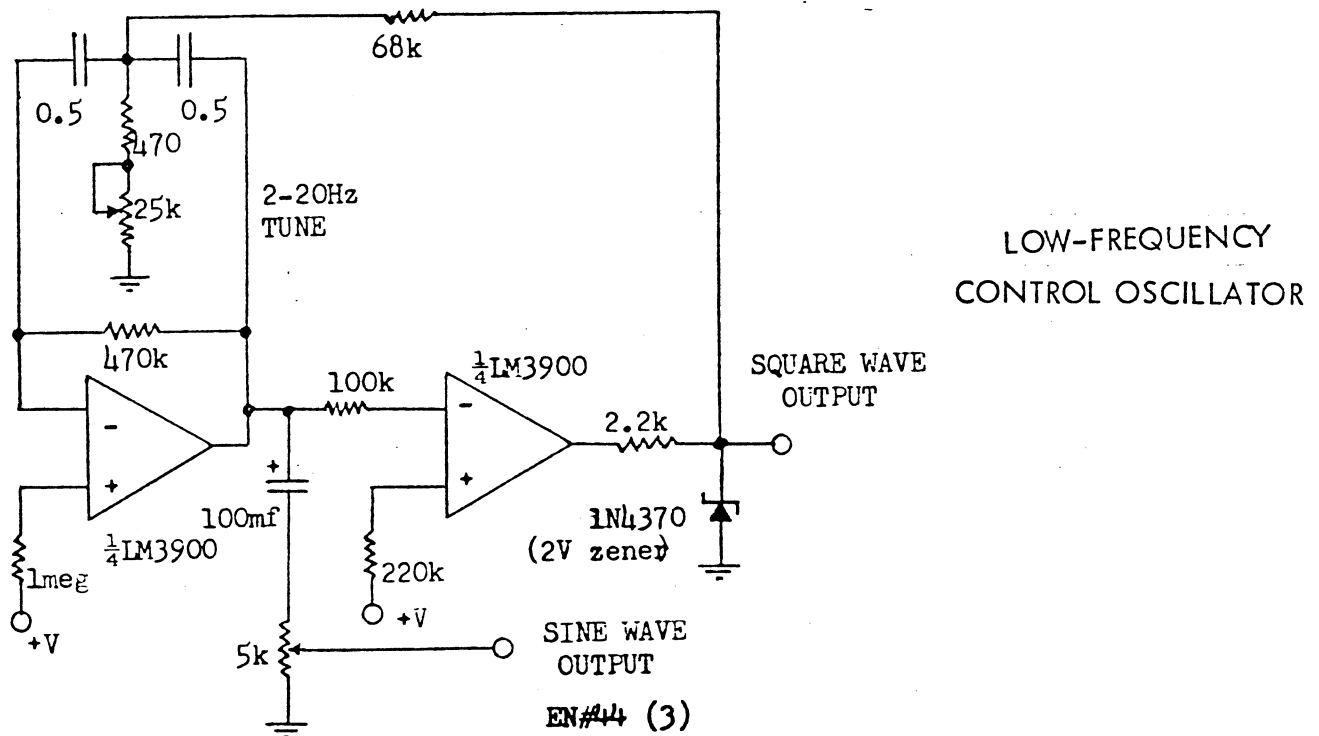


#### 4A. READER'S EQUIPMENT:

Craig Anderton has submitted the following inexpensive function generator based on the Intersil 8038 chip. Craig provides the following description:

The circuit is pretty self-explanatory. The only calibration is to 1) Turn the thing on and let it warm up 15 minutes 2) adjust the frequency control for about 1000 Hz and adjust the "duty cycle" for 50% 3) tune the freq. control to max freq. (bottom setting) and adjust "hi trim" for 20 KHz 4) tune the freq. control to min. freq. (top) and adjust "low trim" for 20 Hz 5) repeat steps 3 & 4 until the freq. control sweeps 20 Hz - 20 KHz.

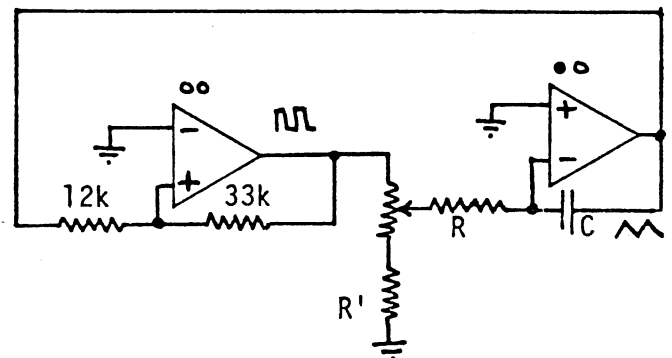
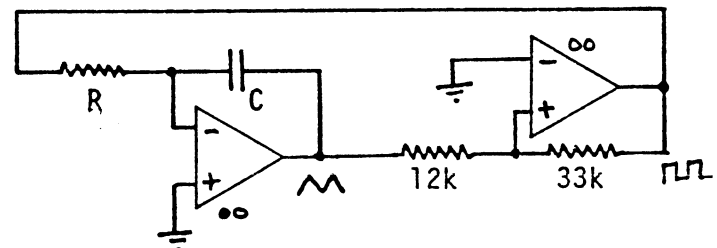




► Q: What's the best circuit for a low-frequency oscillator?

A: Probably most designers use the simple manually tuned triangle-square generator for this purpose. A basic circuit is shown in the figure at the right. A full analysis of this circuit can be found in application note AN-67 (to be published) but it is easy to remember that the frequency of oscillation in this case is not far from  $1/RC$ . The triangle output is at a  $\pm 5$  volt level, and is convenient for vibrato and low frequency control purposes.

Making the oscillator variable is just a matter of putting in a pot as shown in the lower figure at the right. Theoretically there is no limit on the lowest frequencies that can be obtained, but op-amp bias current and a number of other factors may make it total range only about 100:1, which is fine for many purposes. To assure that the oscillator does not stop at the lowest pot setting, some designers add the resistor  $R'$  amounting to about 3% the pots resistance.



Questions Continue on Page 21



## VOLTAGE-CONTROLLED AMPLIFIERS

THE VCA OPTIONS: The main options for the VCA were selected to give good performance and to also be inexpensive. These include one from Dave Rossum based around the CA3046 chip, and one from the ENS-76 series based on an idea of W. Jung. The other two VCA's feature both linear and exponential controls. The first of these is based on a design of Terry Mikulic and uses the CA3080 while the other is a high performance VCA based around the SSM-2020 VCA chip, and was designed by Dave Rossum.

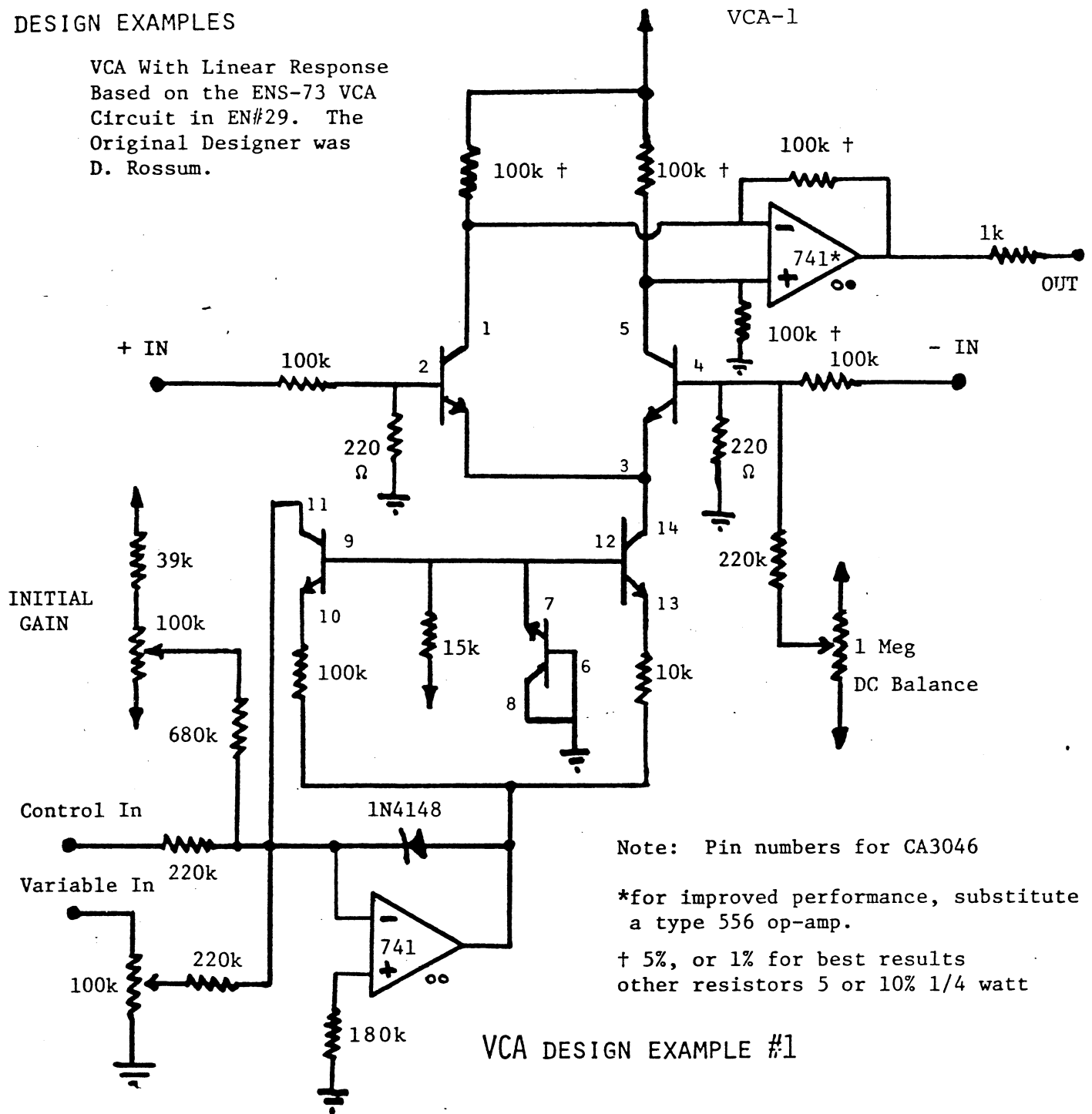
| <u>Preferred<br/>Circuit<br/>Option</u> | <u>Original<br/>Source</u> | <u>Original<br/>Designer</u> | <u>Estimated<br/>Parts<br/>Cost</u> | <u>Notes</u>  |
|-----------------------------------------|----------------------------|------------------------------|-------------------------------------|---------------|
| VCA-1                                   | EN#29 or<br>MEH 5c (9)     | Rossum                       | \$3                                 | ENS-73 Series |
| VCA-2                                   | EN#34 or<br>MEH 5c (10)    | Mikulic                      | \$4                                 |               |
| VCA-3                                   | EN#63                      | Hutchins*                    | \$4                                 | ENS-76 Series |
| VCA-4                                   | EN#87                      | Rossum                       | \$12 (dual)                         |               |

\*includes an idea suggested in several places by W. Jung

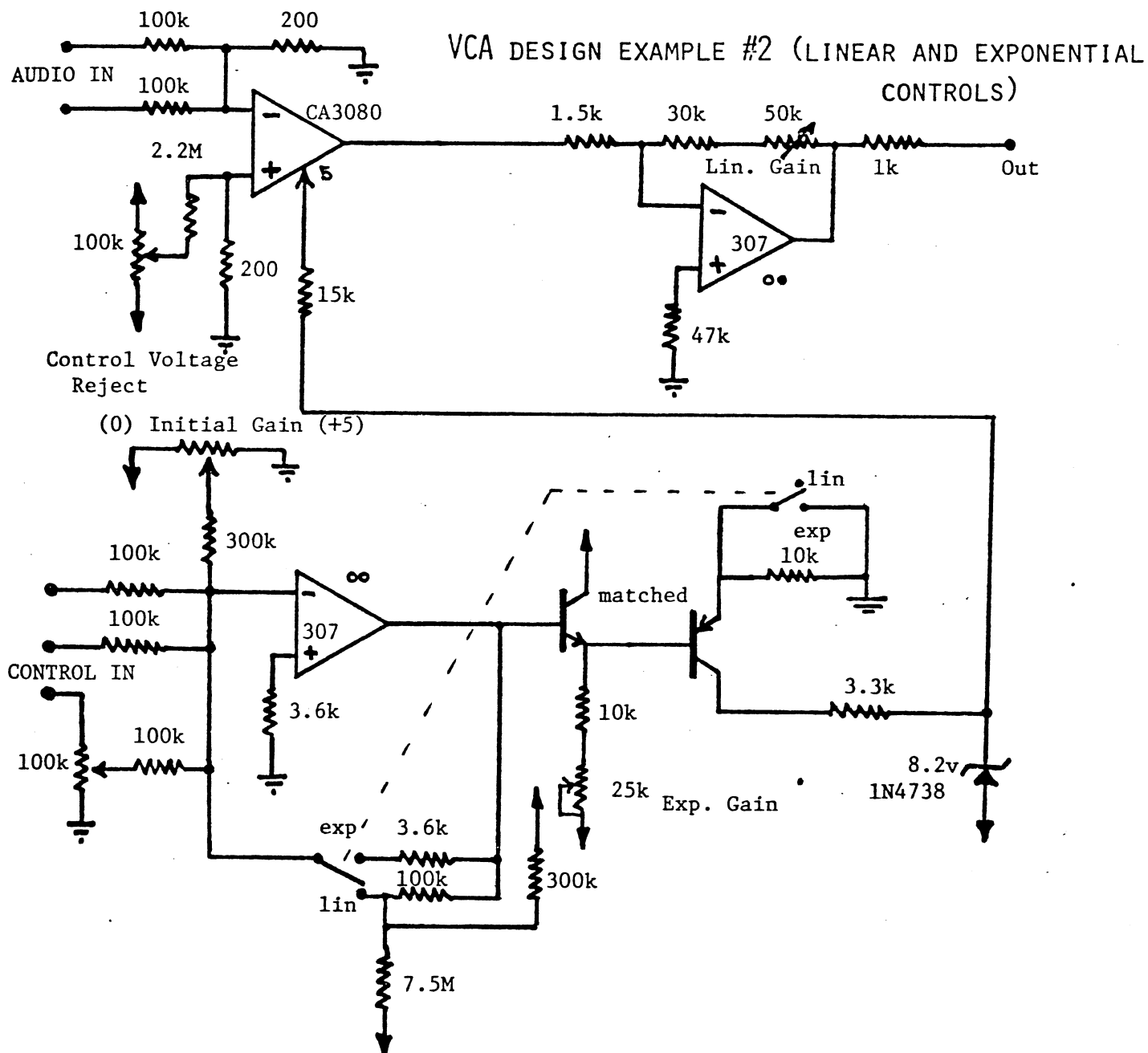


## DESIGN EXAMPLES

VCA With Linear Response  
Based on the ENS-73 VCA  
Circuit in EN#29. The  
Original Designer was  
D. Rossum.



The VCA in the first design example includes the basic two quadrant multiplier that was discussed in detail above, and the linear current source with the refinements mentioned. The circuit has proven very reliable in several applications. Control voltages can sum to +5 to give unity gain. The DC balance trimmer can be adjusted by applying an audio signal to the control input and adjusting for minimum control signal feedthrough in the output. The 1N4148 diode protects the control input from a voltage reversal. The diode connected transistor (pins 6, 7, & 8) is used to drop the base voltage slightly negative. Some designers use a second series diode here, and this can be added if desired. The diode bridge structure could be used as an input attenuator if desired, but for a general purpose VCA, this is not necessary.



The second design example is from EN#34, a circuit designed by T. Mikulic. This circuit may be switched for either linear or exponential response. The VCA has a dynamic range up to 100 db. The linear response is (control voltage/5) and the exponential response is 12db/volt. When the sum of the control voltages is +5, the gain is 1 in either control mode. To adjust, start with the linear mode. Set "initial gain" to "+5" position. Adjust lin. gain so that input and outputs are equal. Set to exp. mode and adjust exp gain for the same level of output. Finally, remove the input signal from the audio inputs and connect to a control input. Adjust the control voltage reject trimmer for minimum feedthrough.



# VCA-3

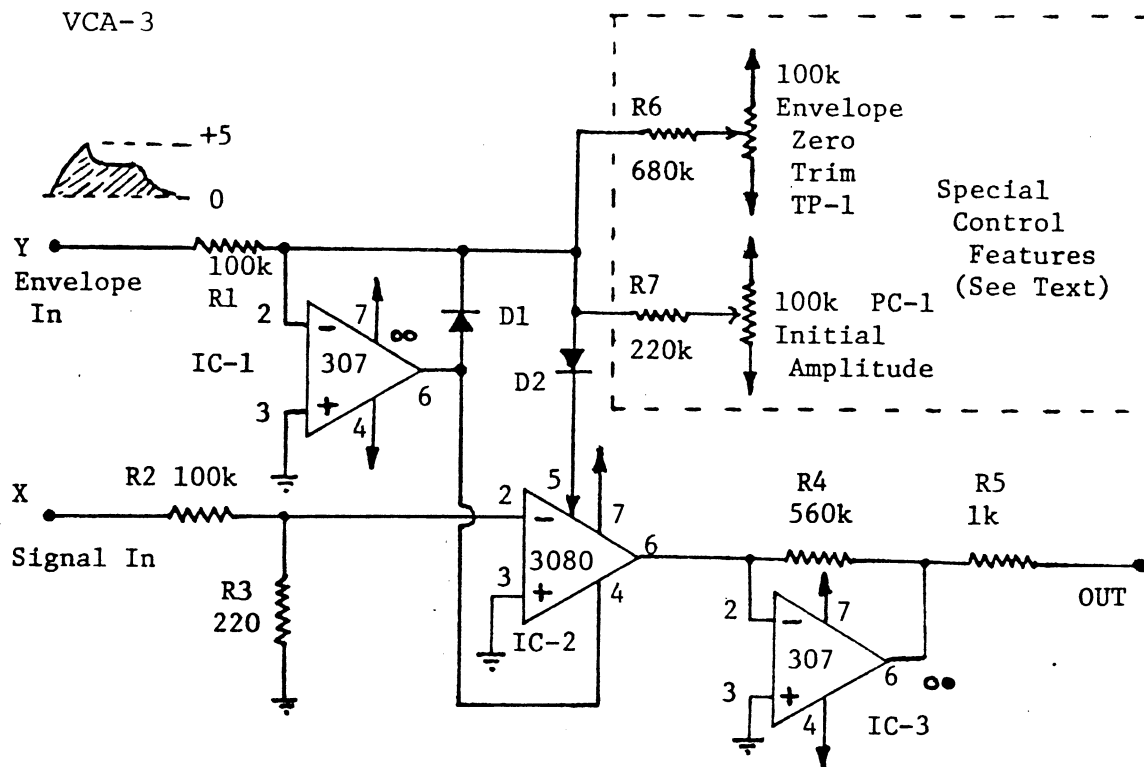


FIG. 5 ENS-76 VCA Option 1

from EN#63

# VCA-4

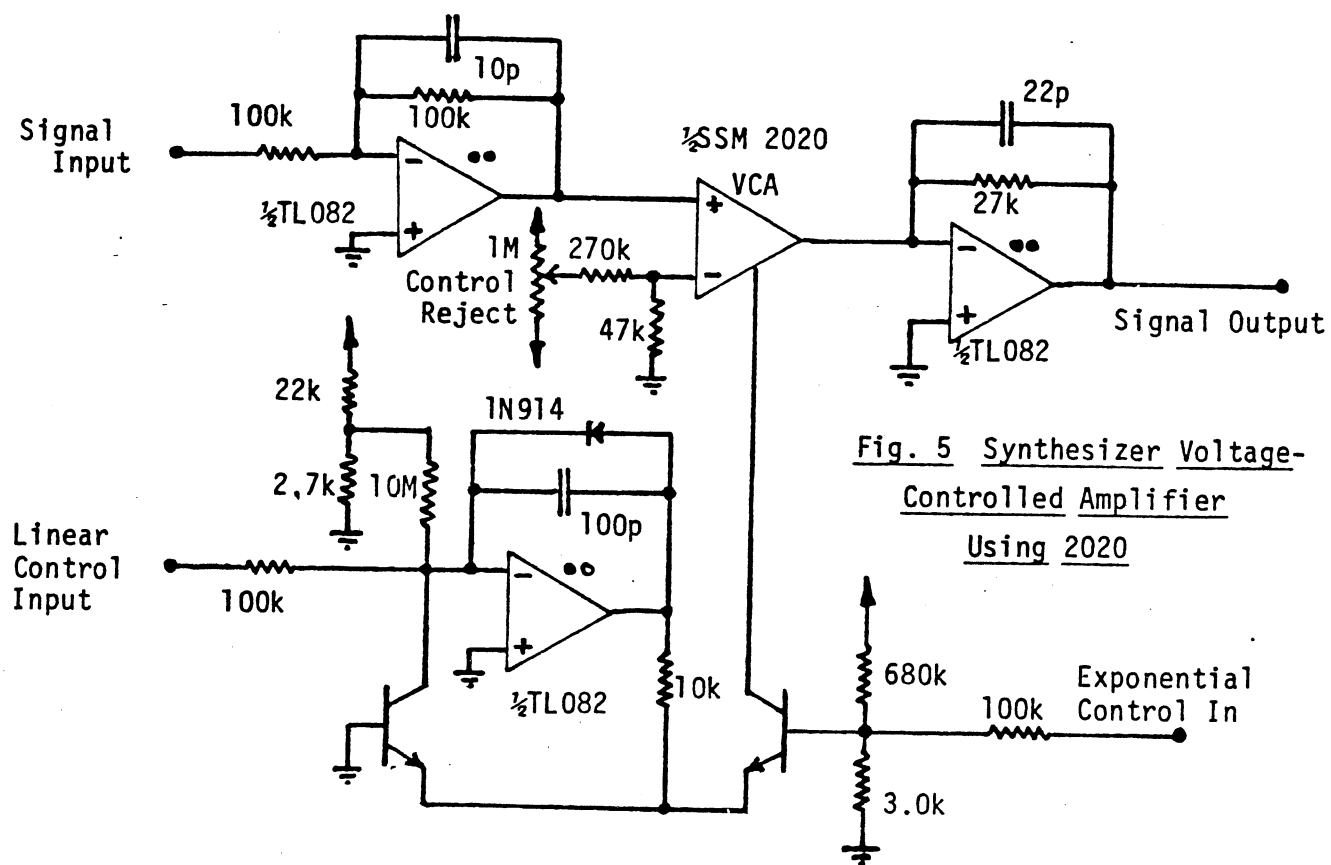


Fig. 5 Synthesizer Voltage-Controlled Amplifier Using 2020

from EN#87



## ENVELOPE FOLLOWERS

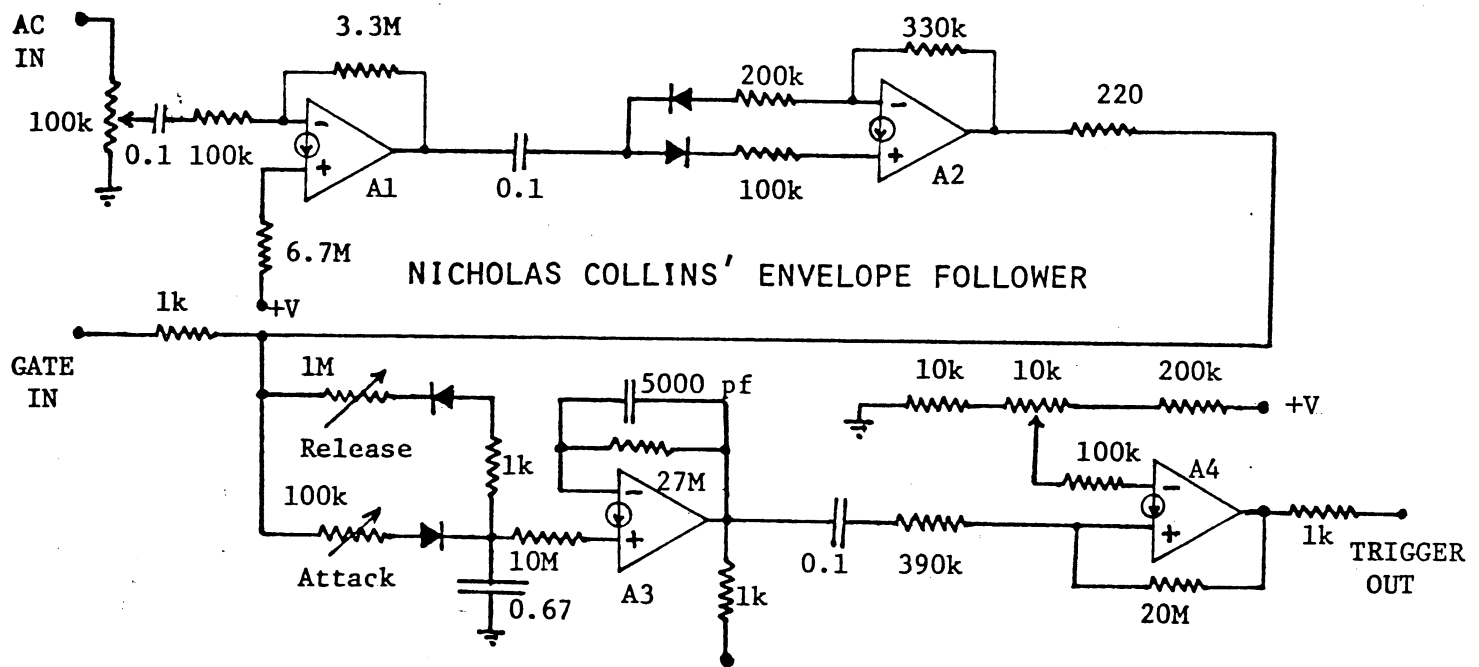
THE ENVELOPE FOLLOWER OPTIONS: Basically, an envelope follower is a device for recovering the amplitude envelope of a signal. It consists of something like a full-wave rectifier, and a low-pass filter. In units intended for a wide range of frequencies, the low-pass filter presents a design problem because it must have a low enough cutoff to remove ripple in the rectified output, and a high enough cutoff to allow the unit to respond to any rapid changes of amplitude. This is not always possible. The first option of this collection is a design of Nicholas Collins and consists of an amplifier, a full-wave rectifier, and another envelope generator, with trigger extraction. It is suitable for generating a controlled envelope from an external input, and to a degree, the output envelope is proportional to the input level. The second option by Robert Iodice is a standard rectifier and filter with added circuitry for recovering a gate and trigger as well. The final option, is a recent circuit from Denny Genovese and has the unusual innovation of using two full-wave rectifiers in cascade, thus multiplying the input frequencies by four so that a low-pass filter with a higher cutoff can be used. This circuit also recovers a gate.

| Preferred<br>Circuit<br>Option | Original<br>Source | Original<br>Designer | Approximate<br>Parts<br>Cost | Notes            |
|--------------------------------|--------------------|----------------------|------------------------------|------------------|
| EF-1                           | EN#60              | Collins              | \$4                          | + Gate, Trigger  |
| EF-2                           | EN#86              | Iodice               | \$3                          | + Gate, Trigger  |
| EF-3                           | EN#88              | Genovese             | \$4                          | + Gate           |
| EF-4                           | EN#89              | Hutchins             | \$9                          | ripple reduction |



The circuit below which was submitted by Nicholas Collins does not avoid the pitfalls described above, but it does add a degree of control that could well give a substantial improvement in some areas. This circuit has as its heart an actual envelope generator which serves as the "low-pass filter." The attack and release controls of the generator adjust the time constants of the "filter" so that increasing and decreasing voltages are handled differently. If you like, you can think of this circuit as an AR envelope generator which is driven by a funny sort of "gate." This "gate" is the output of the full wave rectifier, and it varies in peak amplitude. The ripple of this "gate" is filtered by the low-pass action of the AR generator.

Amplifier A1 is just a signal amplifier, while A2 is a full-wave rectifier. A3 is a buffer for the AR generator setup. Finally A4 serves as a differentiator which produces a trigger whenever the envelope changes rapidly. This permits the unit to be used to trigger other envelope generators from a live signal for example. As can be seen, the entire unit is built from a single LM3900 quad CDA, making it inexpensive and compact. Note that an actual gate can be applied to the circuit to make it serve as an AR type of envelope generator when not in use as an envelope follower.



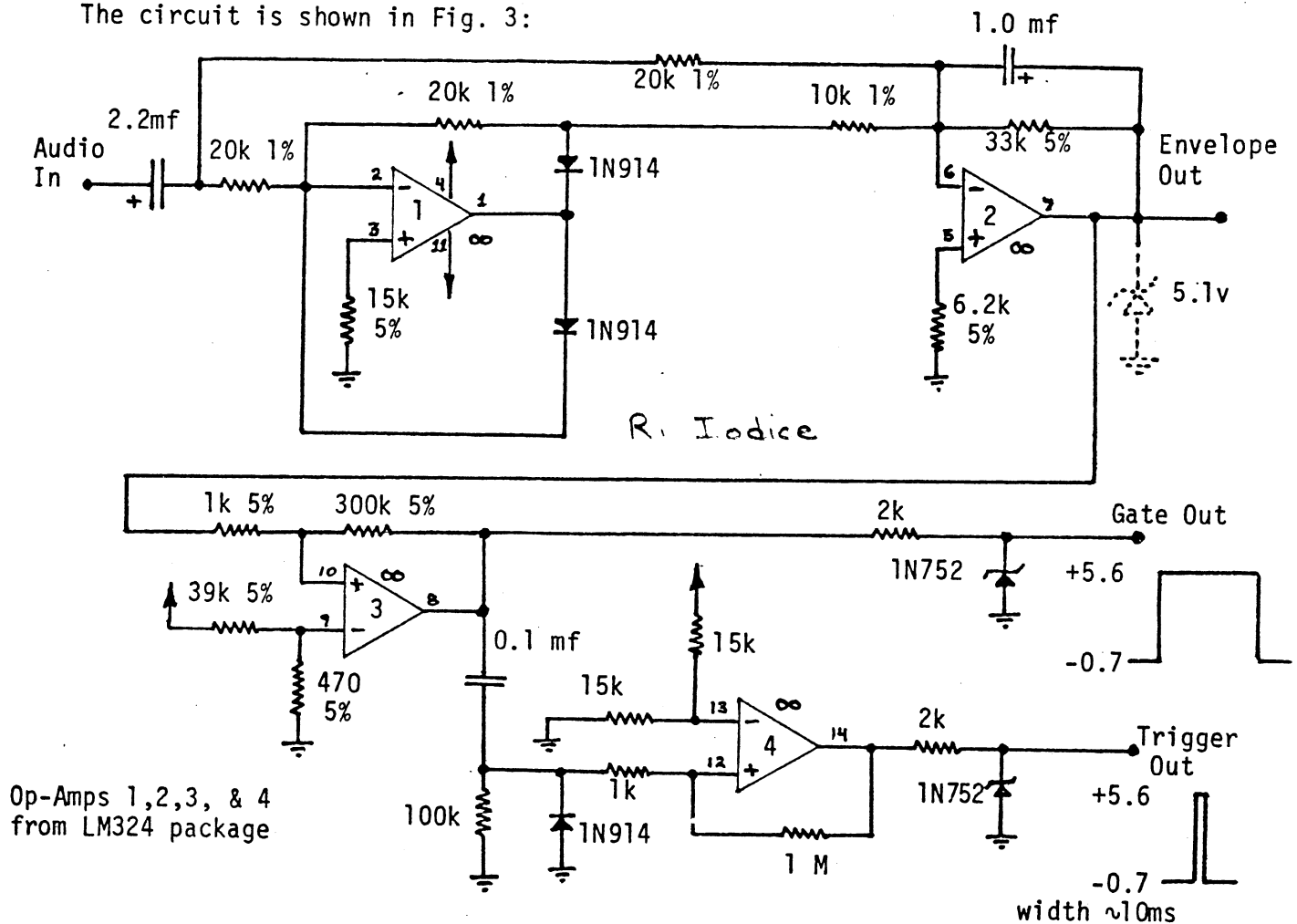
A1, A2, A3, A4  
each 1/4 LM3900

ENVELOPE OUT

ENVELOPE OUT

Power supply +9 to +18

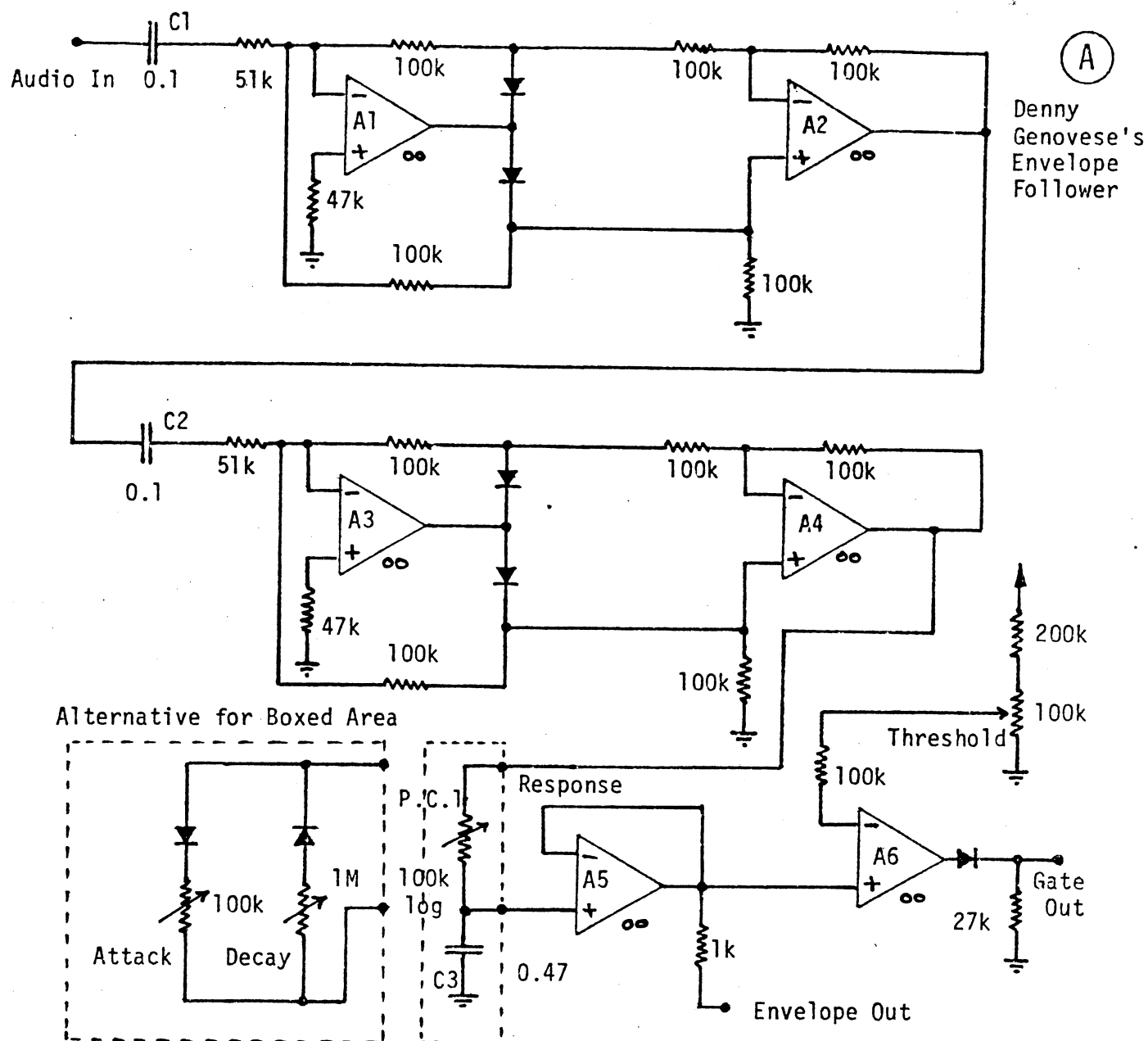
The circuit is shown in Fig. 3:



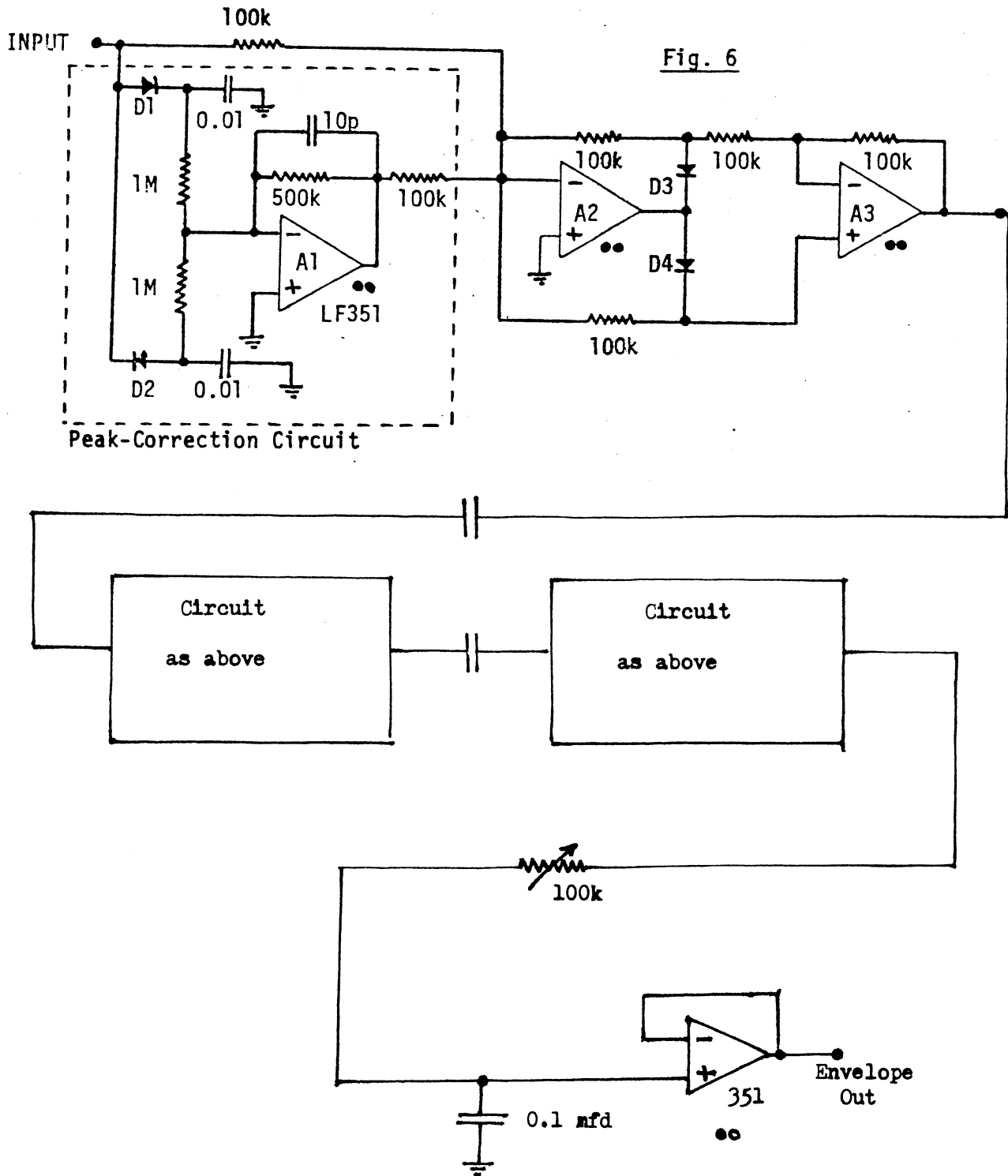
The 33k feedback resistor on op-amp 2 sets the gain of the envelope extractor. In this case a 1Vp-p sine wave yields a 500 mV D.C. output. A 5.1 volt Zener such as a 1N751 may be used at the envelope extractor output to prevent outputs greater than ~5 volts since the ENS-76 VCA requires no more than +5V in for full output.

The circuit works well with audio inputs from 40 Hz to about 10 kHz. At 80 Hz (~low 'E' string on a guitar) the ripple is 5% with the 1.0 mfd filter capacitor.

EN#86 (21)



The circuit below is derived from information in EN#89 but is otherwise unpublished in complete form.





## VOLTAGE-CONTROLLED FILTERS

VCF's: We will begin our coverage of VCF's with a reprint in its entirety of a VCF design article with two state-variable VCF's as the final product. We will then look at a standard four-pole filter in two different forms, one based on the CA3080, and the other on the SSM2040 IC. Next we will look at a generalization of the four-pole, the polygon filter, and at a variable-slope filter. Finally, we will take a brief look at a couple of very new filters. The VCF options are:

| <u>OPTION</u> | <u>SOURCE</u> | <u>DESIGNER</u> | <u>APPROX. COST</u> | <u>NOTES</u>          |
|---------------|---------------|-----------------|---------------------|-----------------------|
| VCF-1         | EN#71         | Hutchins*       | \$9                 | State-Variable        |
| VCF-2         | EN#71         | Hutchins*       | \$12                | S-V with VC Q         |
| VCF-3         | EN#41         | Hutchins        | \$10                | four-pole (CA3080)    |
| VCF-4         | EN#87         | Rossum          | \$15                | four-pole (SSM2040)   |
| VCF-5         | EN#97         | Hutchins        | \$27                | 4,6, & 8 pole polygon |
| VCF-6         | EN#72         | Hutchins        | \$11                | variable-slope        |
| VCF-7         | EN#90         | Hutchins        | \$25                | high-ripple           |
| VCF-8         | EN#92         | Hall            | \$17                | switching capacitor   |

In this installment, we will be giving two of the VCF designs for the ENS-76 series. The first of these is a new formulation of the state-variable filter, and has been made as simple as possible. The second filter is a revision of the filter given in EN#37. We expect to give two more filters in the future. One of these to come will have a cascaded state-variable configuration for 24db slopes, and the other will have a voltage-variable slope. The reader should keep in mind that many of the features in these filters are "interchangable" in the sense that they may be used in structures other than the one they are actually presented in. The reader should feel free to select features from the different options to arrive at exactly the filter he needs.

## INTRODUCTION TO THE NEW VCF DESIGNS

The VCF designs we will be using are all state-variable. We have found that this filter can not be improved upon either as a general purpose filter or as a special purpose one. We have changed around a few structures here and there in previous designs, and used a few new parts, but basically there is one major improvement which we have implemented, and this is the control of the runaway Q at high frequency.

We should point out right away that the compensation technique for controlling Q which we give here is not original with us. Actually, we discovered it in a recent paper [Sergio Franco, "Use Transconductance Amplifiers to Make Programmable Active Filters," Electronic Design, Sept. 13, 1976, pg. 98]. Ian Fritz pointed out to me that this same general technique was discussed by R. Sparkes and A. Sedra in "Programmable Active Filters," IEEE J. Solid State Circuits, Feb. 1973. This paper in turn leads us back to an earlier application of a similar technique to the biquad circuit in "The Biquad: Part I - Some Practical Design Considerations," by Lee Thomas, IEEE Trans. Circuit Theory, CT-18, May 1971 [Also reprinted in Active Inductorless Filters edited by S. K. Mitra, IEEE Press (1971)].

So what is it? Basically it is just a matter of adding in a couple of capacitors on the order of a few picofarads to produce a phase lead, compensating for phase shift at high frequency, thus stabilizing the filter at high Q and high frequency. Our previous VCF designs showed instabilities at high frequency and high Q. The circuit in EN#30 which used type 595 multipliers was very bad in this respect. It was found that the capacitor C9 in this case did stabilize the circuit. However, this design was then abandoned in favor of one using the CA3080 as a control element. The final result was the filter in EN#37. It turns out that it takes only a few picofarads to stabilize this filter. Since most stray capacitances are on this order, it is not too surprising that trouble with this filter did not appear for all builders, and certainly not for all builders at the same limiting points. None the less, it is a good idea to add in these capacitors or at least check to make sure that the response is not just marginally stable.

The basic structure of the state-variable filter is shown in Fig. 1. This should be familiar by now. It is a summer and two integrators, and the structure is stabilized by negative feedback from both integrators. As Q increases, the negative feedback from the first integrator decreases. Thus, higher Q is associated with a system that is less stable than lower Q situations. There is another way that negative feedback can be reduced. This is indicated by considering a more realistic model of the state-variable as shown in Fig. 2. Here we show in addition to the summer and integrators, a stage which adds some phase lag. This is not something we have added to the circuitry, but something that was there all the time. It is due to phase shifts across the IC's which is due to capacitances between the elements inside the IC's. This phase shift increases with higher frequencies. As this phase shift increases, negative feedback is lost in favor of positive feedback. Clearly if the phase shift is 180°, it is all positive feedback. One solution to this problem is shown in Fig. 3 where a phase lead network has been added to the inputs of the integrators. We add this only to the integrators because they each include two IC's while the summer is only one IC. Also, it is really only necessary to make some correction, and

not to correct for every little detail of the phase response. In fact, there are probably several break points in the phase response of the integrators while the network shown to provide the phase lead only corrects for the largest of these, or perhaps just gives an average which takes care of the problem. An exact correction would require accurate modeling of all the active devices, which would be unnecessary for our purposes.

We should note that the value of the capacitor  $C_{qc}$  is just a few picofarads when the resistors shown are the standard 100k and 220 ohm attenuating resistors used on the inputs of CA3080's. A capacitance of say 3pf is difficult to work with. We found in working with the circuit of EN#37 that when we soldered jumpers of about 1 inch onto the 100k resistors (so that we could easily solder on capacitors) that the two jumpers of about 1" separated by about 1/4" made a substantial improvement to the stability. For this reason, we suggest using attenuators for the CA3080's of 10K and 22 ohms for these circuits. This means that the compensating capacitors can be ten times as large. We will see that with better quality op-amps, even this change means that the compensation may be only 10pf or so. But the important thing is that the scheme does work.

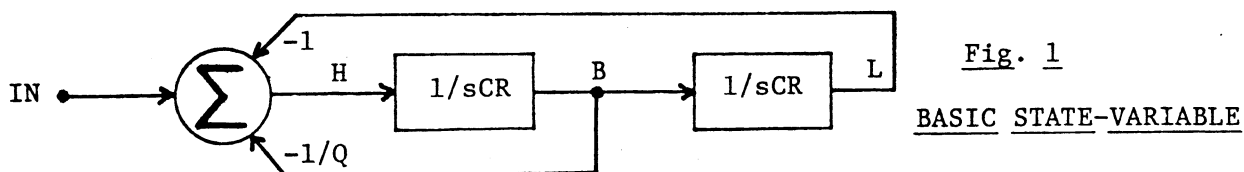


Fig. 2 STATE-VARIABLE WITH EXISTING PHASE SHIFTS

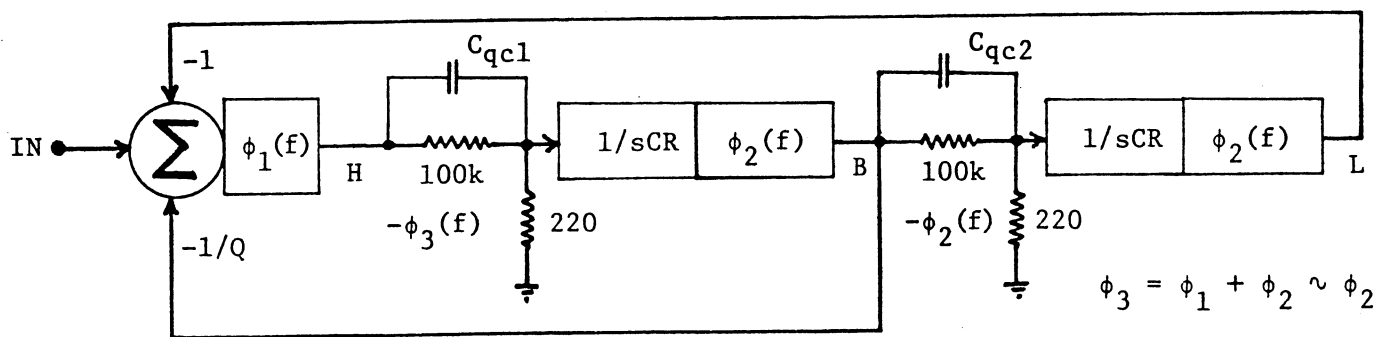
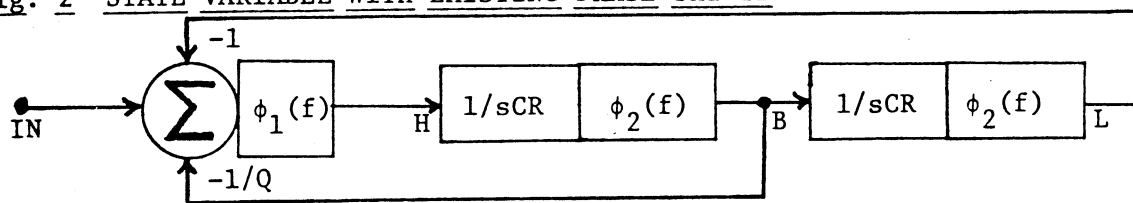


Fig. 3 STATE-VARIABLE WITH PHASE SHIFT COMPENSATION

## ENS-76 VOLTAGE-CONTROLLED FILTER - OPTION 1

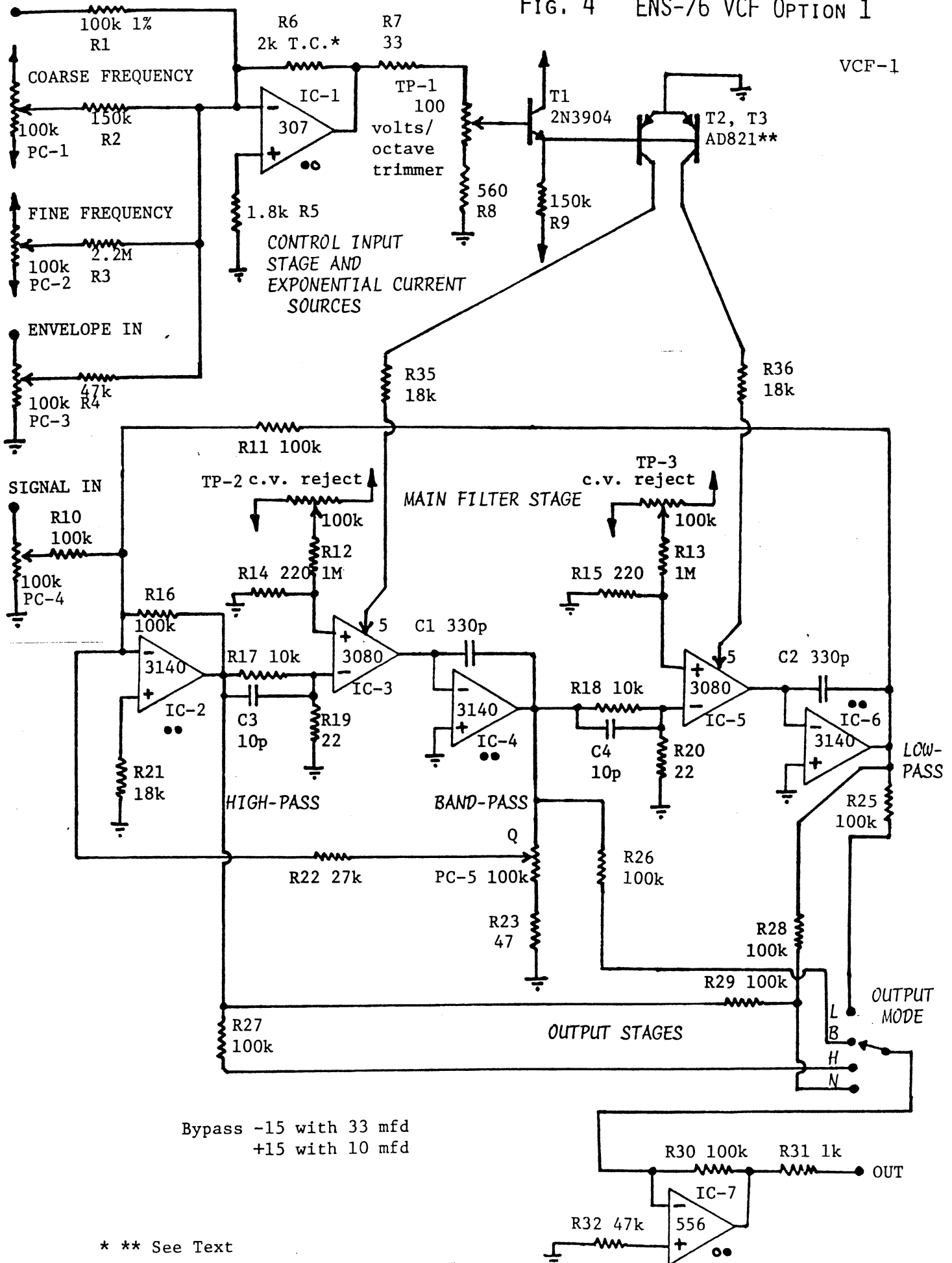
The first of the ENS-76 VCF options is shown on page 16. This filter features a state-variable structure giving second order responses for low-pass, band-pass, and high-pass. A manual control for Q is provided, and a notch output is also available. The filter has a range from below 1/10 Hz up to about 18 kHz, and Q is variable from about 1/4 to well over 500.

The circuitry of the filter is quite similar to what we have done before. There are about four changes which we should point out. First, there is a somewhat different exponential current source, which is a dual source. Secondly, we use the new RCA CA3140 for the filter op-amps because it seems to have a very small phase shift relative

1 volt/oct In

FIG. 4 ENS-76 VCF OPTION 1

VCF-1



to other op-amps we tried. Thirdly, phase-lead circuitry has been used to prevent Q-enhancement. Finally, we used a mode switch on the output instead of using separate output jacks for each of the modes. We will discuss these changes in more detail below.

The exponential current stage is really just the basic structure described by Terry Mikulic in EN#37. A standard summing network (IC-1) sums the control voltages and scales them to about 20mv change for a one volt change of the input control. This is trimmed to about 18mv by the 100 ohm trimmer TP-1. T1 is an emitter follower which sets the  $I_o$  current for the exponential converter ( $I_o = 15/R9$ ) and also drives the converter transistor T2. A second transistor T3 is driven in parallel with T2 and thus provides an identical current (assuming the transistors T2 and T3 are matched). Resistor R6 is a 2k +3500ppm/°C temperature compensating resistor (Tel Labs type Q81). For non-critical work, you can use a 2k 5% carbon resistor. Likewise, T2-T3 is specified as a matched pair (Analog Devices AD821) but for non-critical work, you can use a pair of 2N3906's glued together. These exponential currents are driven through resistors R35 and R36 into the control pins (pin 5) of the CA3080's. Since the collectors of T2 and T3 can reach no higher than about zero volts, the current into pin 5 of the CA3080's is limited to less than 15volts/18k = 0.83ma. The resistors R35 and R36 serve to limit this current, and thus limit the upper frequency of the filter to about 18 kHz, but otherwise have no effect on the circuit and can be ignored during analysis. In previous circuits, we have driven such a pair of resistors from a single exponential current source. For currents such that a substantial voltage is developed across these resistors, the resistors tend to divide the current equally between the two CA3080's. However, for low currents, there is very little voltage drop across these resistors and the current may not be shared equally. Thus, the dual source in this example is probably a better idea.

In a number of lab tests, we found that the new RCA CA3140 was about an order of magnitude better than the LM307 when used in state-variable filters. This was determined when we were testing for Q-enhancement. Since we wanted a very high-Q filter, we decided to use the CA3140 here and deal with the fact that it is not a satisfactory output driver by using an output buffer (IC-7). The low bias current of the CA3140 also permits the filter to operate at very low frequency. We actually found that we could ring the filter as low as 0.002 Hz, but at that low a frequency, we did not have enough time to make any measurements of control accuracy. Thus, we are just claiming response down to 1/10 Hz as we have verified this.

Before going on to describe how the phase correction circuitry is adjusted, we should say a few words about how the filter frequency is controlled by the control current, how the Q is determined, and how the control voltage reject trim pots are adjusted. First we note that the basic equation for the CA3080 is:

$$I_{out} = 19.2 \cdot I_{ABC} \cdot V_{di}$$

where  $I_{ABC}$  is the control current into pin 5, and  $V_{di}$  is the differential input voltage (the actual voltage difference between pins 2 and 3 of the CA3080). Since there is an attenuator of 10,000:22 on the input, and the (+) terminal is effectively zeroed, the input voltage ( $V_{in}$ ) is related to  $V_{di}$  by  $V_{in} = 456 \cdot V_{di}$ . Substituting this back into the basic CA3080 equation, we get:

$$R_{eq} = V_{in}/I_{out} = 23.7/I_{ABC}$$

where  $R_{eq}$  is the equivalent resistance of the CA3080 ( $R_{eq} = 23.7k$  when  $I_{ABC} = 1$  ma). Consulting Fig. 5 on page 6 of this issue, we see that the center frequency of the filter is  $1/2\pi R_{eq}C$ . We saw above that the current into pin 5 of the CA3080's on page 16 was limited to about 0.83ma, corresponding to  $R_{eq} = 28.6k$ , and since  $C = 330pf$ , we get an upper frequency limit of about 17 kHz, corresponding well to the 18 kHz observed. We can also obtain from the information on Fig. 5 on page 6 that Q is equal to the inversion of the gain from the bandpass output back to the high-pass output. With the Q control in its minimum position (away from R23), the gain is  $R16/R22 = 100k/27k = 3.7$  corresponding

to a Q of 0.27. With the Q control in its maximum position (toward R23), the gain is  $(47/100k) \cdot (100k/27k) = 0.00174$  corresponding to a Q of 574. For higher Q, R23 may be set to 22 ohms with some possible loss of stability margin. The control voltage rejection pots are adjusted so that when the coarse frequency control is adjusted through its full range, the deflection at the low-pass output is a minimum. Set both at their midpoints to start with, then alternately adjust them until the deflection is a minimum. For non-critical work, you can just ground the (+) terminal of both 3080's and leave out TP-2, TP-3 and associated resistors R12-R15.

The phase lead circuitry was installed according to the general principles outlined above. We will describe here the exact method of determining the proper value for the compensation capacitor, and this will serve to indicate how the value should be set in other filters and for any exact testing by individual builders. The first step is to set a medium value of Q in the range of 5 to 50. It is not essential that the exact theoretical value for Q be known, but it is helpful to know this. In this circuit, we removed R22 and put a 1 meg resistor from the band-pass output to the (-) input of IC-2. This meant that the theoretical Q should have been 10. Next, the Q is measured as a function of frequency. At some frequency, it will be noted that Q has risen somewhat above the low frequency value (See Fig. 5). The Q at the highest frequency the filter will be used at should be noted. During testing, this was 25 kHz for the present filter. The rise of Q at higher frequencies is known as Q-enhancement. Once this enhancement is detected, it should be corrected. This is a matter of installing capacitors C3 and C4. We started with 18pf for these, and the values of Q were again measured. Note that this resulted in overcompensation since the Q comes down at higher frequencies. However, excellent stability was obtained with the 18pf value. We then tried 10pf for these capacitors and found it leveled off the Q quite well. In testing for higher values of Q, we found we could get values of Q up to 1000 at 10kHz with the 10pf capacitors. However, at times, such a high Q at 25 kHz resulted in instability. Rather than increase the values of C3 and C4, we thought it better to cut back the maximum frequency by increasing R35 and R36 to 18k.

Since Q is in the range of 5 to 50 for this case, a number of methods can be used to measure Q, and these can be found in the first report on bandpass filters in this newsletter. Note that the ringing method can be used even if a scope with an accurate time base is not available. All that is required is a VCO or other function generator that produces a sawtooth of fairly well known frequency. The setup is shown in Fig. 6. The sweep frequency is adjusted until the amplitude at the end of the trace on the scope is  $1/e = 37\%$  of its starting value. Denoting this frequency by  $f_s$ , it is clear that the ring time of the filter is  $1/f_s$ , and  $Q = \pi f_0 T_{ring} = \pi f_0 / f_s$  where  $f_0$  is the center frequency which has previously been determined by measurement with a sinewave input. Another useful method of leveling off the Q would be to control both the VCO and the VCF from the same voltage (assuming they track) as shown by the dotted line in Fig. 6. With this setup, the trace on the scope has a constant envelope if the Q is constant with frequency. If the end of the trace gets smaller as frequency goes up, less compensation is needed. If the end of the trace rises with frequency, more compensation is needed. This is probably one of the fastest ways of getting satisfactory leveling of Q. It might also be interesting to try a somewhat larger capacitor in the stage driven by the high-pass (C3) and a slightly smaller value for C4. The reason for this can be seen by consulting Fig. 3 where we see that the capacitor C3 compensates for two phase shifts while the capacitor C4 compensates for only one. However, it works well with both capacitors the same.

Finally, we decided to use an output mode switch because it will save some space on the panel, and otherwise we would need separate output buffers for each stage [the CA3140 is not satisfactory - see EN#69 (13)]. We thus use a buffer such as the type 556 since it is fast enough and is a satisfactory output driver. In any event, this additional op-amp would have been needed for a notch output. The inversion of IC-7 is also useful since a positive going transition on the input will cause the BP output to start to ring in the positive direction first, and this seems the most natural.

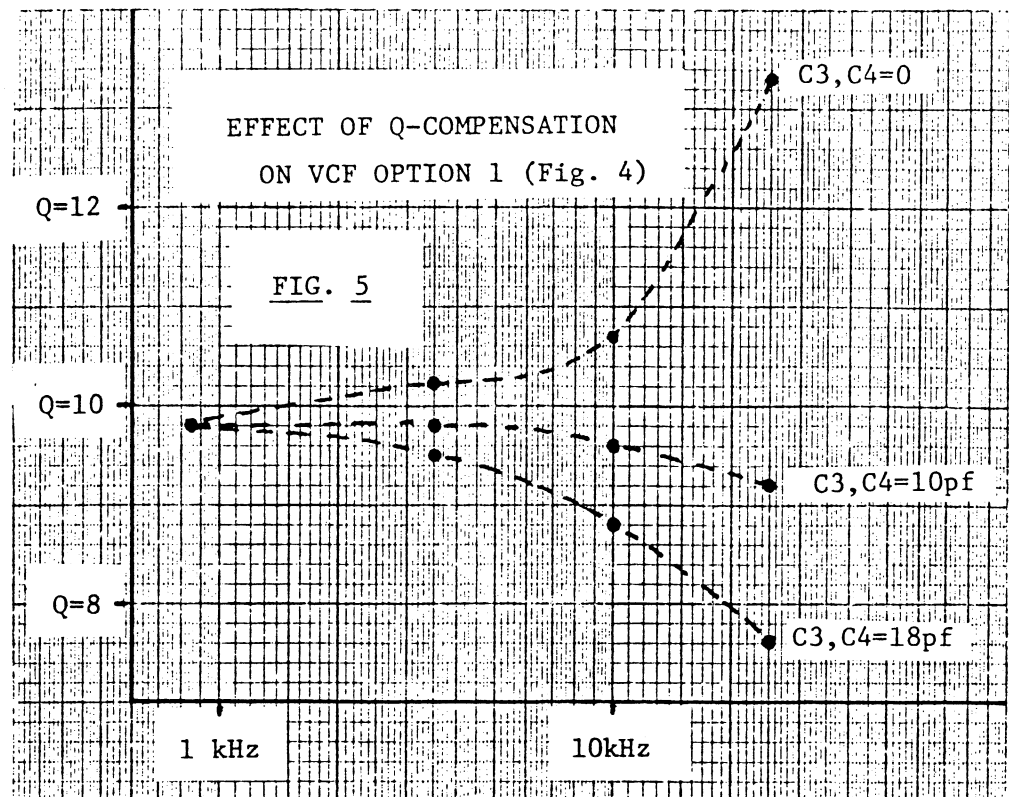
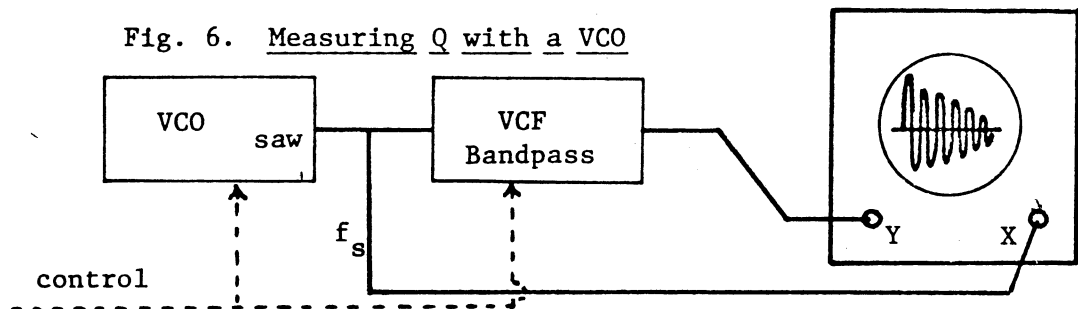


Fig. 6. Measuring Q with a VCO



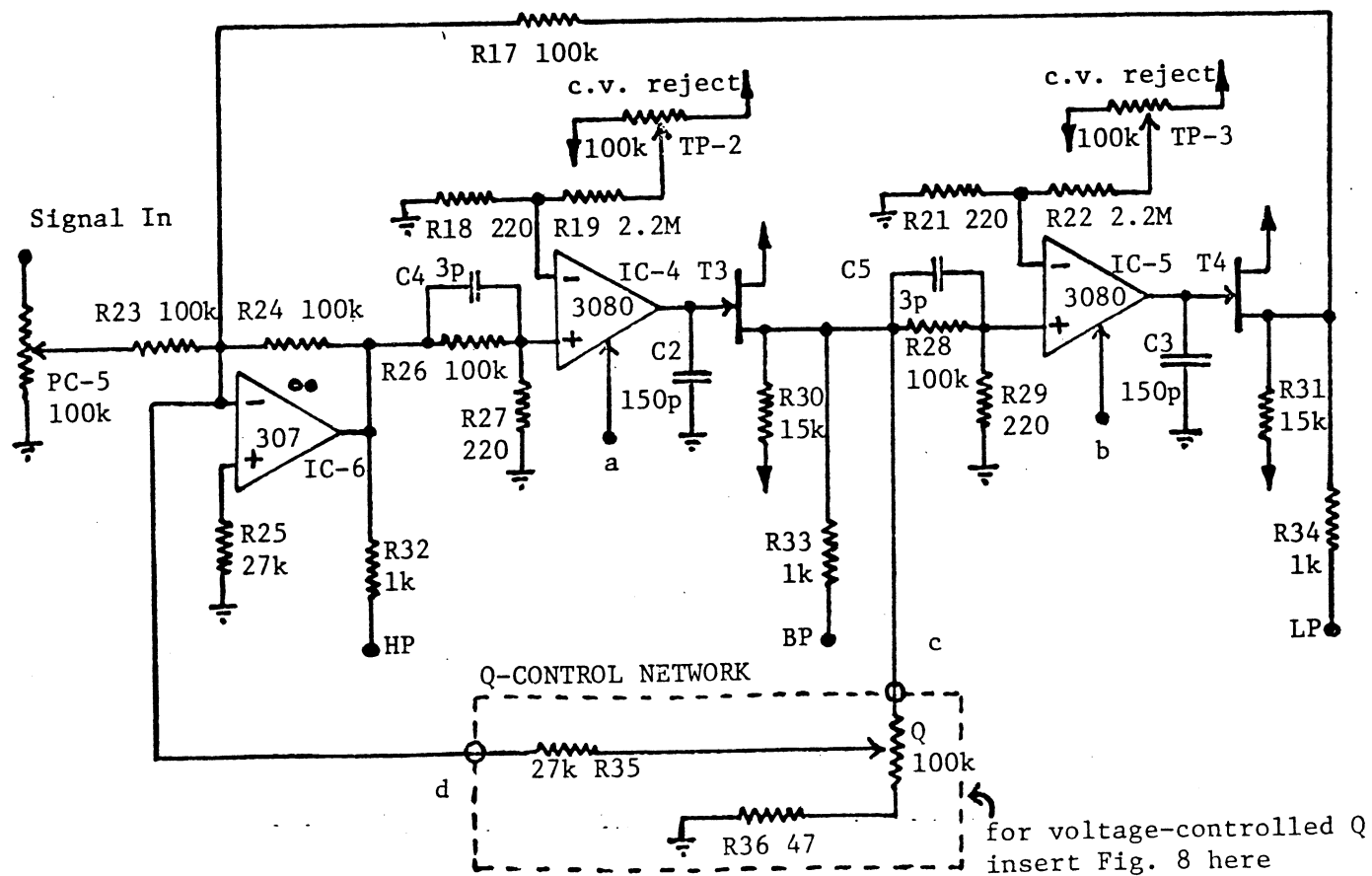
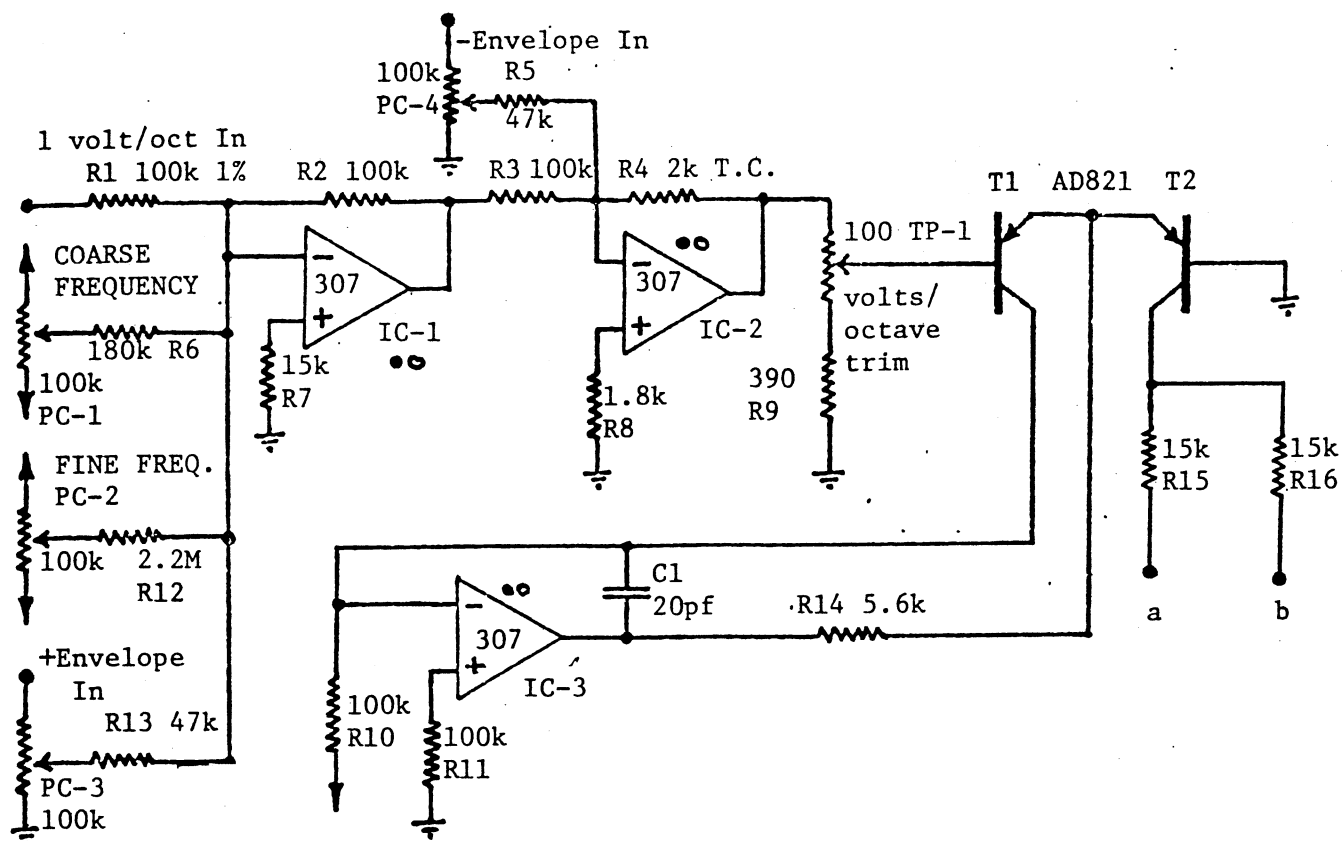
## ENS-76 VOLTAGE-CONTROLLED FILTER - OPTION 2

The second option for the ENS-76 VCF is shown in Fig. 7 and Fig. 8. This filter is a modification of the EN#37 design. The exponential converter section of this filter uses a single current source split by two resistors ( $R_{15}$  and  $R_{16}$ ) to supply the two CA3080's. The exponential source from Option 1 could be used here. One advantage of the current source in Option 2 is that it has inputs for both positive and negative excursions of frequency in response to a positive going voltage. Also, if linear modulation of filter frequency should become as popular as linear FM in VCO's, this current stage can be given linear control by supplying additional current to IC-3's summing node.

The actual filter is similar to Option 1 except it uses FET buffers instead of op-amps, and has parallel outputs for all three filter functions, High-Pass, Band-Pass, and Low-Pass. We show in Fig. 7 a manual Q control network to keep the diagram simple, but the voltage-controlled Q section is shown in Fig. 8. This voltage-controlled Q section could also be used with Option 1. The voltage-controlled Q section is somewhat different from the EN#37 design. We use here cascaded inverting summers to sum the Q-control voltage (IC-6 and IC-7). T5 and T6 form an exponential current source that controls the CA3080, IC-8. When the control sum is zero (output of IC-7 is zero), the current into pin 5 of IC-8 is just  $I_0 = 15/R_{39} = 15/150k = 0.1ma$ . The equivalent

FIG. 7 ENS-76 VCF OPTION 2

VCF-2



for voltage-controlled Q  
insert Fig. 8 here



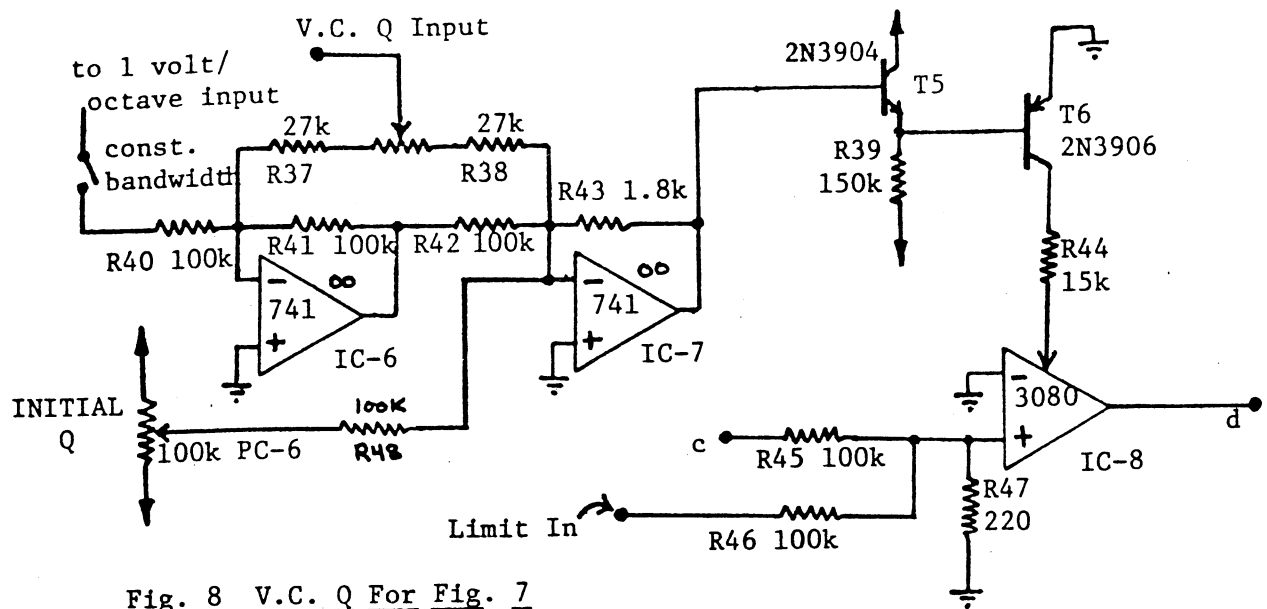
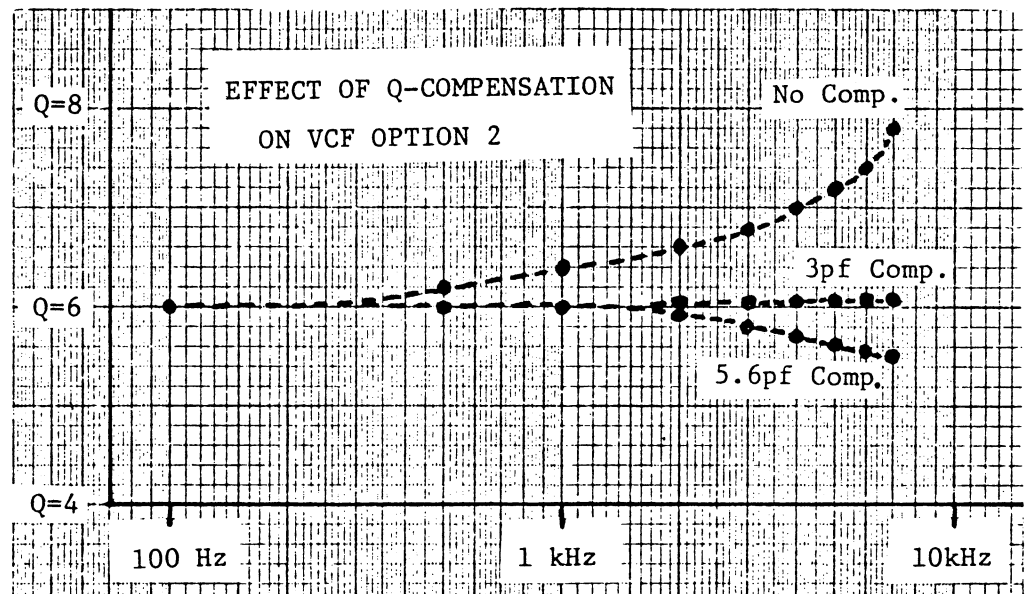


Fig. 8 V.C. Q For Fig. 7

Fig. 9

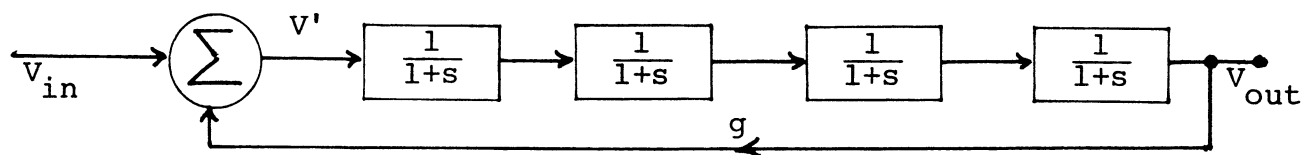


resistance of the CA3080 is thus  $23.7/0.1\text{ma} = 237\text{k}$ . Thus, the gain from BP back to HP is  $100\text{k}/237\text{k} = 0.42$  corresponding to a Q of  $1/0.42 = 2.4$  or so. Now, we provide a switch to connect R40 to the 1 volt/octave control voltage for the filter's frequency control section. A one volt change there will produce a 2:1 change of Q, and thus keep the bandwidth of the filter constant (resulting in a constant ring time independent of frequency). A variable v.c. Q input is provided as well. Note that to shut this one off the pot must be accurately centered, or the controlling source can be removed. The arrangement as shown will allow a unipolar envelope to either adjust the Q up or down depending on the setting of the V.C. Q Input control PC-7.

The filter is compensated for Q enhancement much as Option 1 was. The graph of Q and frequency as a function of the compensating capacitor is shown in Fig. 9. Note that while the optimum value (3pf) seems smaller than the 10pf used in Option 1, since the attenuating resistor is 100k in Option 2 while it was only 10k in Option 1, the 3pf here is equivalent to 30pf for a 10k attenuator. In fact, persons building this circuit new should make R26 and R28 10k, and R27 and R29 equal to 22 ohms, and then use about 30 pf to compensate for Q (thus 30pf for C4 and C5). It is easier to work with a value of 30pf than it is to work with 3 pf since 3 pf is on the order of normal stray capacitances. The tests in Fig. 9 show the result with voltage-controlled Q, not with the manual Q.

\* \* \* \* \*

VCF-3 and VCF-4 are four-pole filters with regeneration. The basic structure of these filters is shown below:



The  $1/(1+s)$  units are just voltage-controlled first-order filters, and the relationship between  $V'$  and  $V_{out}$  is thus:

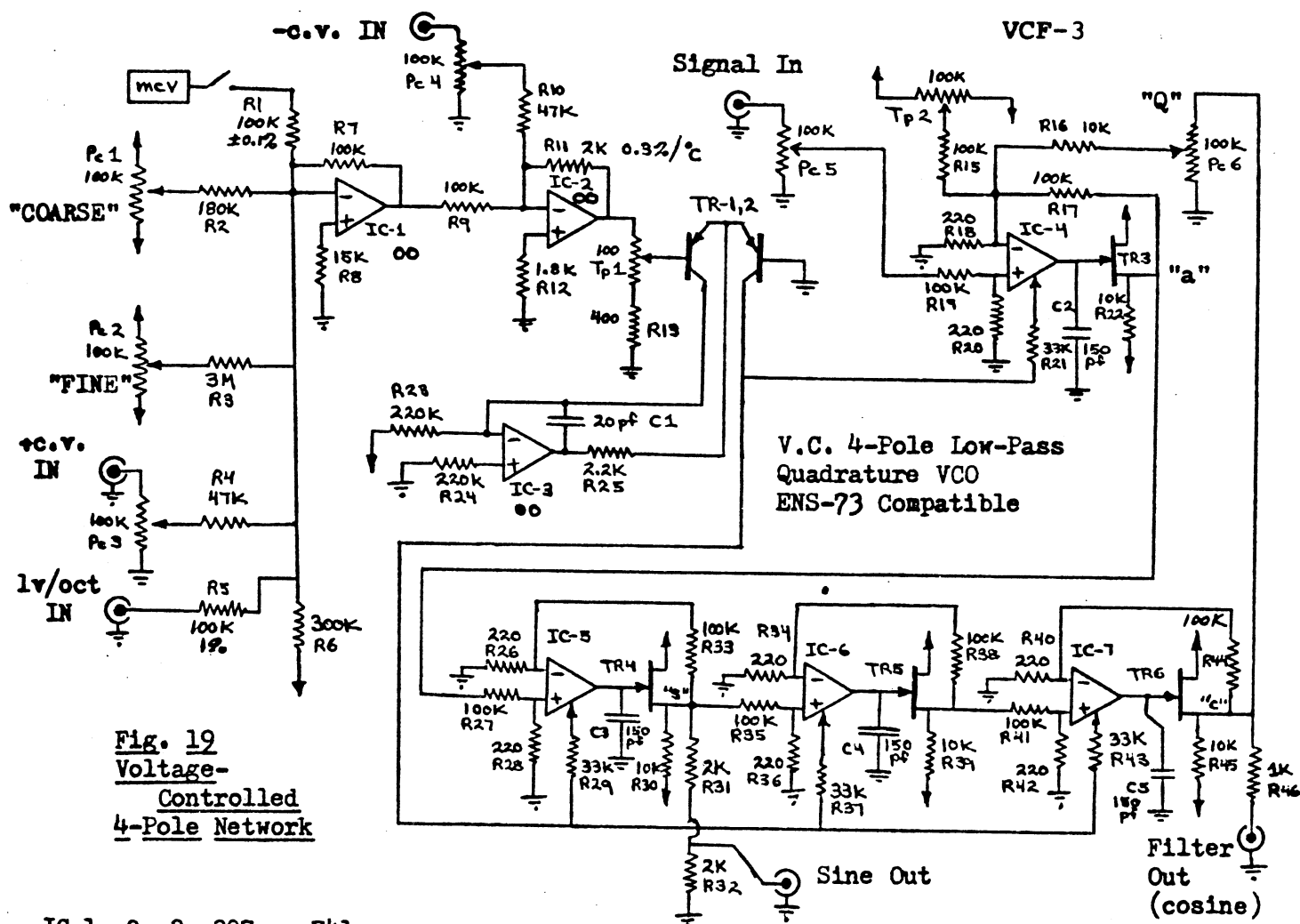
$$V_{out}/V' = 1/(1+s)^4$$

The relationship between  $V_{in}$  and  $V_{out}$  is then:

$$V_{out}/V_{in} = 1/[(1+s)^4 - g]$$

A full analysis of the second equation will show that as  $g$  increases, the frequency response will begin to peak at the corner frequency, and this means that a relatively poor corner that results from four first-order sections in cascade is vastly improved ("corner peaking"). When  $g=4$ , the circuit oscillates producing a sine wave at its corner frequency. This is generally limited by a slight clipping on the first stage, but since this is followed by three stages of low-pass filtering, the sine wave at the output can be excellent.

\* \* \* \* \*



**Fig. 19**  
**Voltage-**  
**Controlled**  
**4-Pole Network**

IC-1, 2, 3: 307 or 741

IC-4, 5, 6, 7: CA3080

TR-1,2: Matched Pair AD821

TR-3, 4, 5, 6: 2N3819 FET's

Adjust  $T_{p1}$  for 1 volt/octave

Adjust  $T_{p2}$  for symmetric clip at "a" during oscillation at low frequency.

The 2k 0.3% temp. comp. resistor and TR-1,2 should be in close thermal contact.

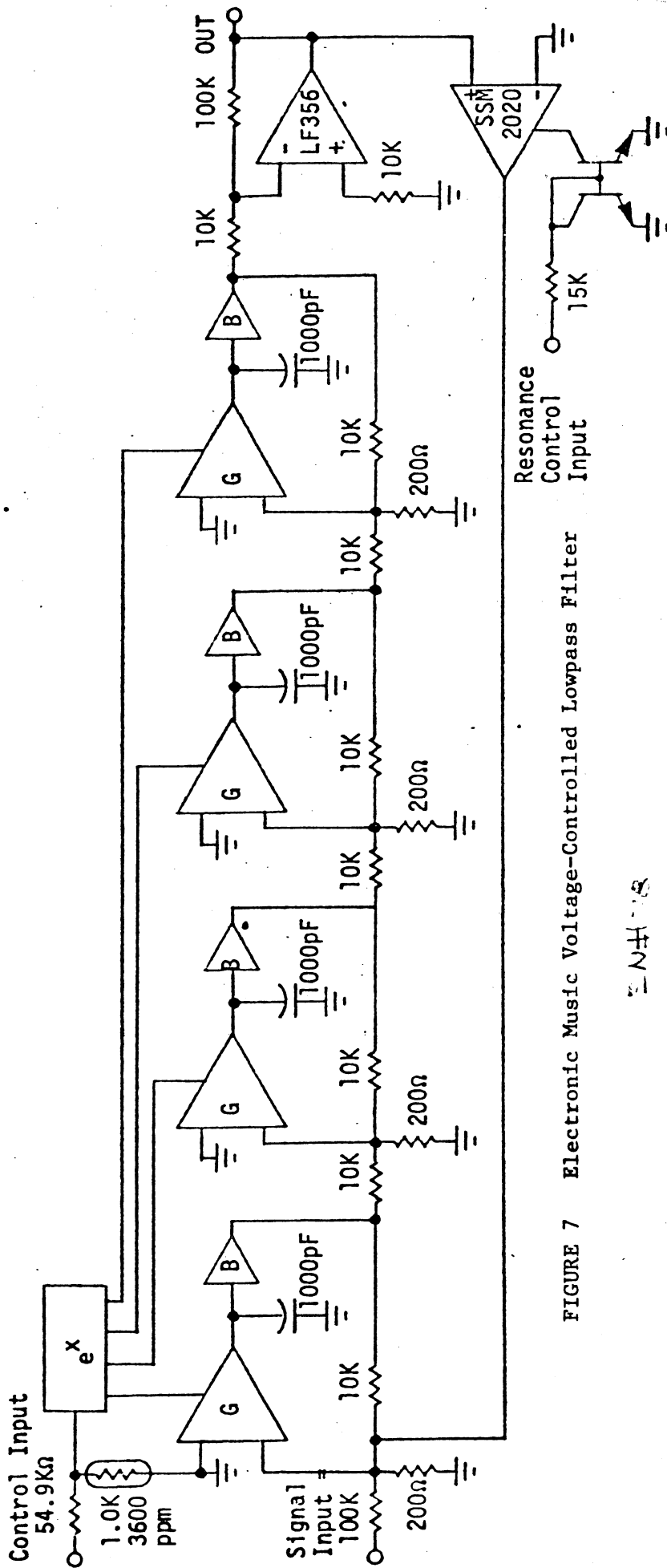
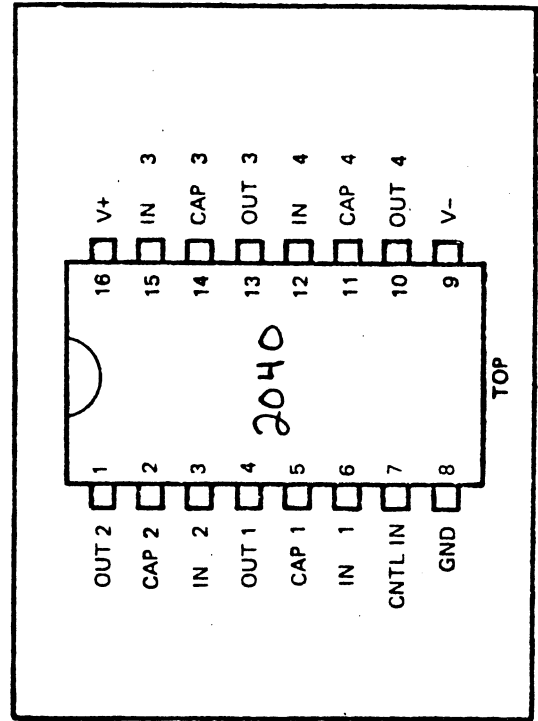


FIGURE 7 Electronic Music Voltage-Controlled Lowpass Filter

2040



Filters VCF-5 through VCF-8

Four additional VCF designs are given in this collection. For full details, see the original newsletters indicated by the page number at the bottom of the schematics.

VCF-5: This is a variable slope VCF, the slope can be varied smoothly from 6db/oct. to 36 db/octave.

VCF-6: This is a special purpose high-ripple, 6db, 6th order Chebyshev filter.

VCF-7: This is a special switching capacitor or "commutating" type of filter.

VCF-8: This design, based on two SSM-2040 chips can be used for fourth, sixth, or eighth order, and is a "polygon" filter, a generalization of the fourth-order low-pass. In fourth order, it is the same as a fourth order low pass, as in VCF-3 and VCF-4.

\* \* \* \* \*

## THE ENS-76 HOME BUILT SYNTHESIZER SYSTEM - PART 6:

- by Bernie Hutchins

### ENS-76 VCF OPTION 3, VARIABLE-SLOPE FILTER:

This third option for a VCF is a special type of VCF which we have not presented in the past. This filter works on a variable slope principle (See report starting on page 3 of this issue) and a discussion of the specific variable slope scheme begins on page 10. The reader should review the theory at some time, but for now he need just realize that basically we have to construct a series cascade of exponentially controlled first order low-pass sections. The first order section we will use has been discussed in various places (see EN#58 for example) and we need say no more about it here. It consists of a CA3080 and an op-amp integrator of which A7 and A13 are typical in Fig. 2 on the next page. Since we have to build six exponential converters, we can't afford to make them too fancy. Here we use fixed attenuators to set one volt/octave, and use unmatched 2N3904-2N3906 pairs for the exponential converting transistors. A typical exponential converter stage can be seen as A1, T1, and T7 in Fig. 2. We just construct six of these virtually identical stages, cascade them, and add appropriate driving voltages.

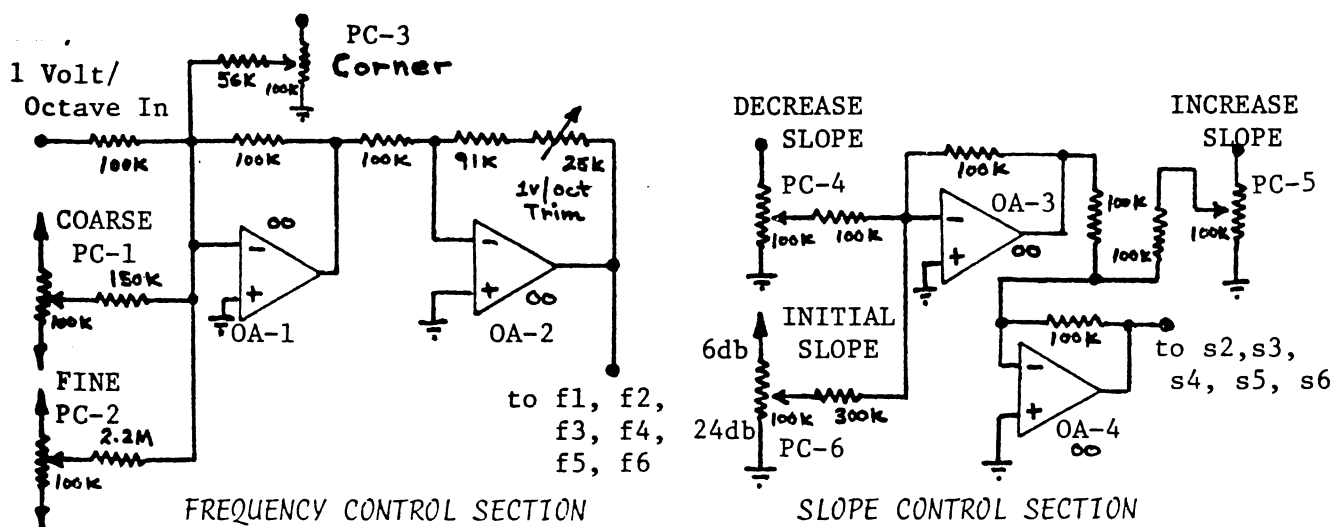
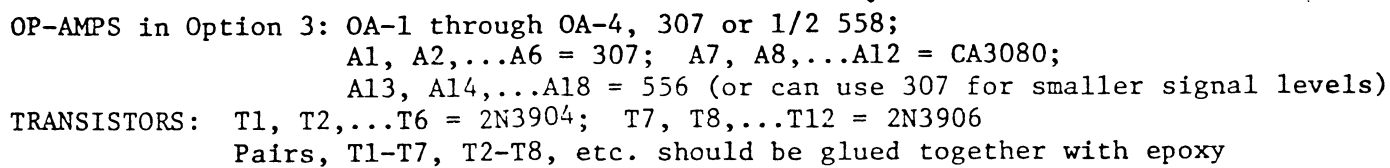
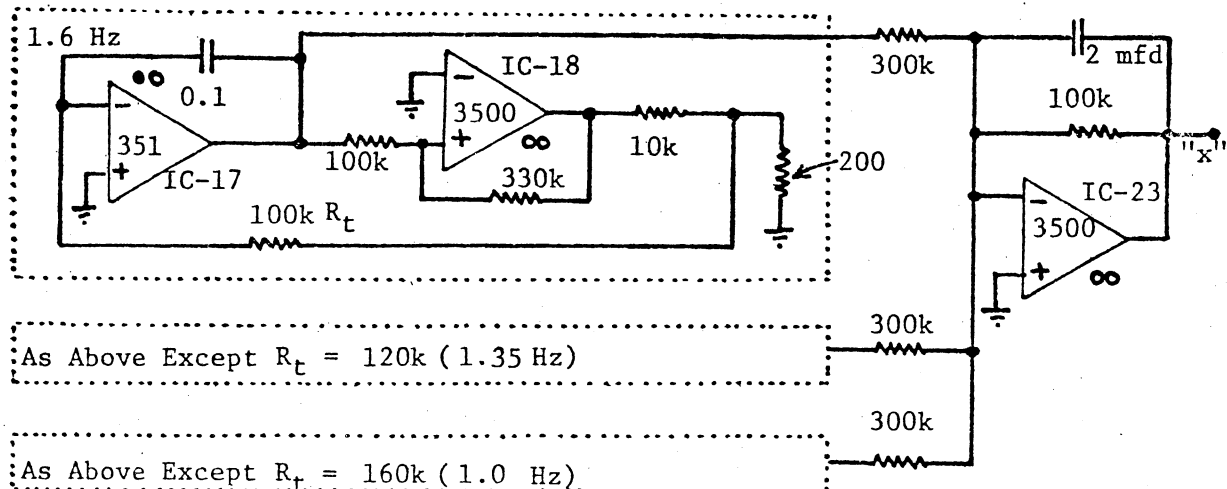
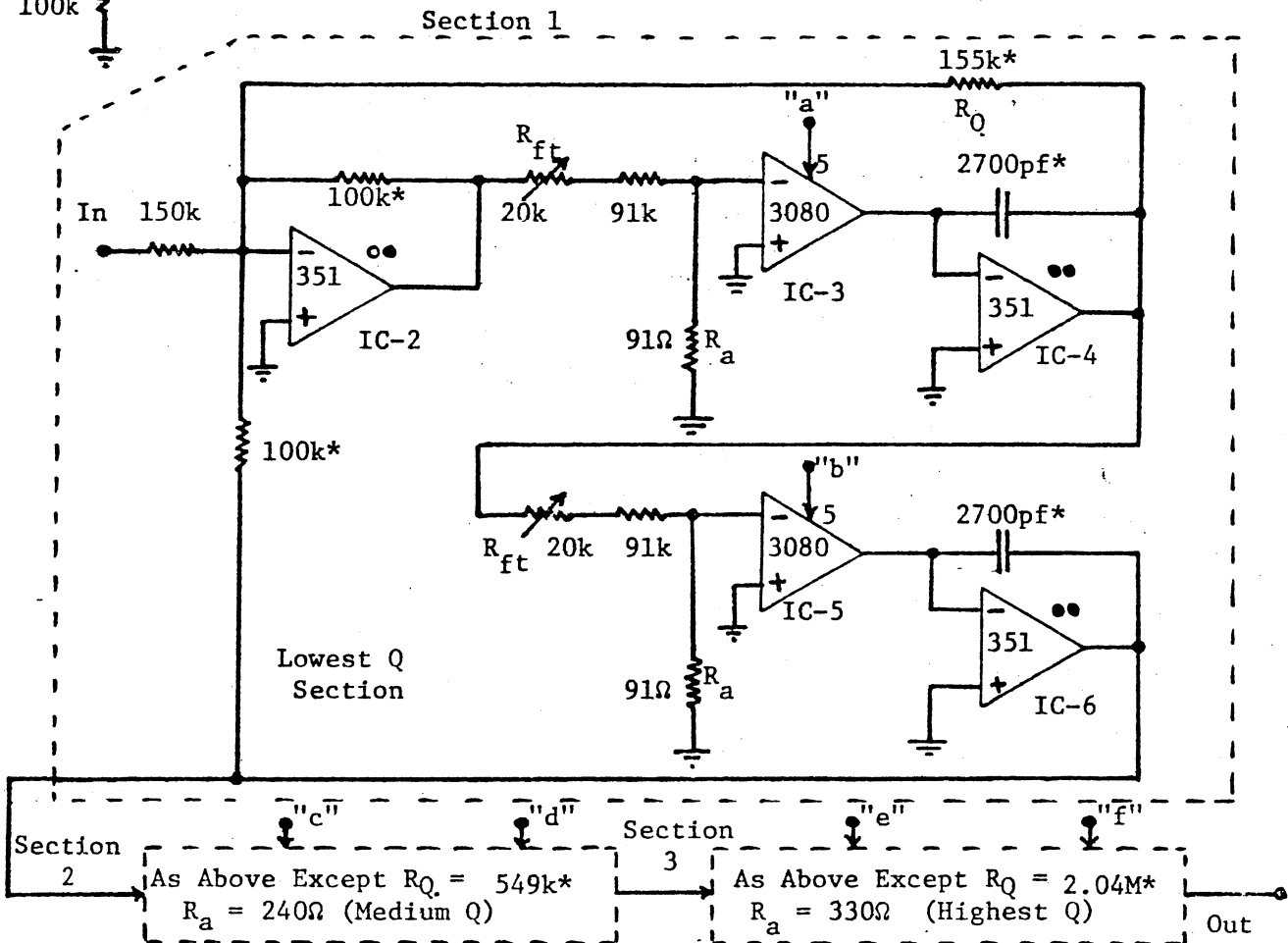
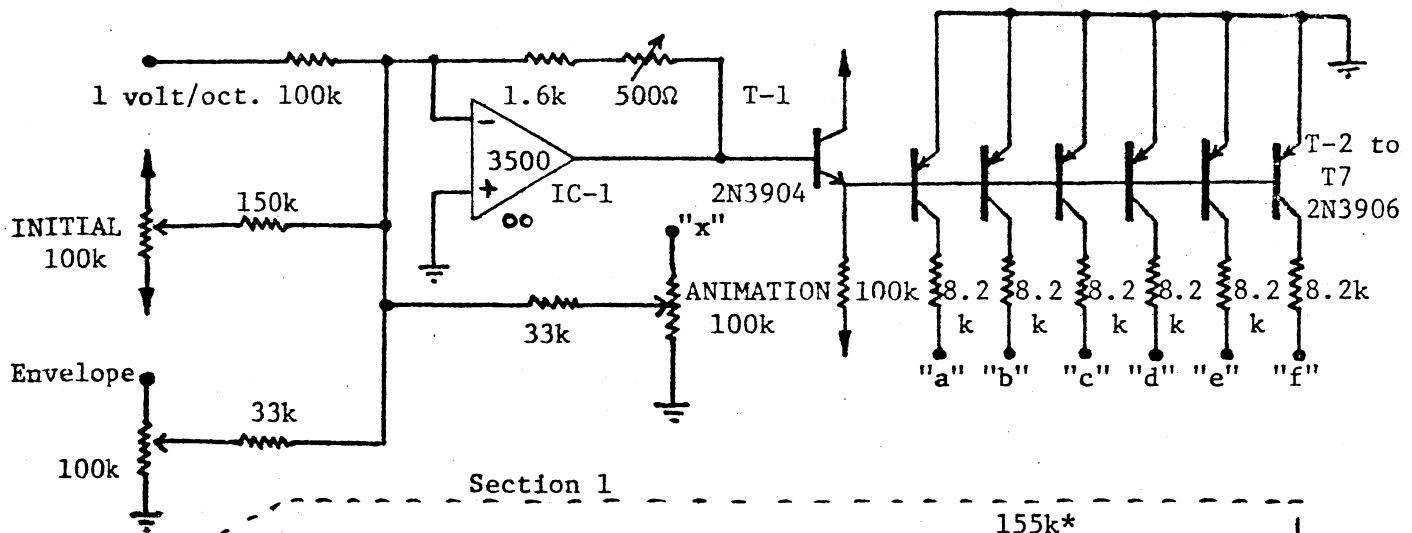
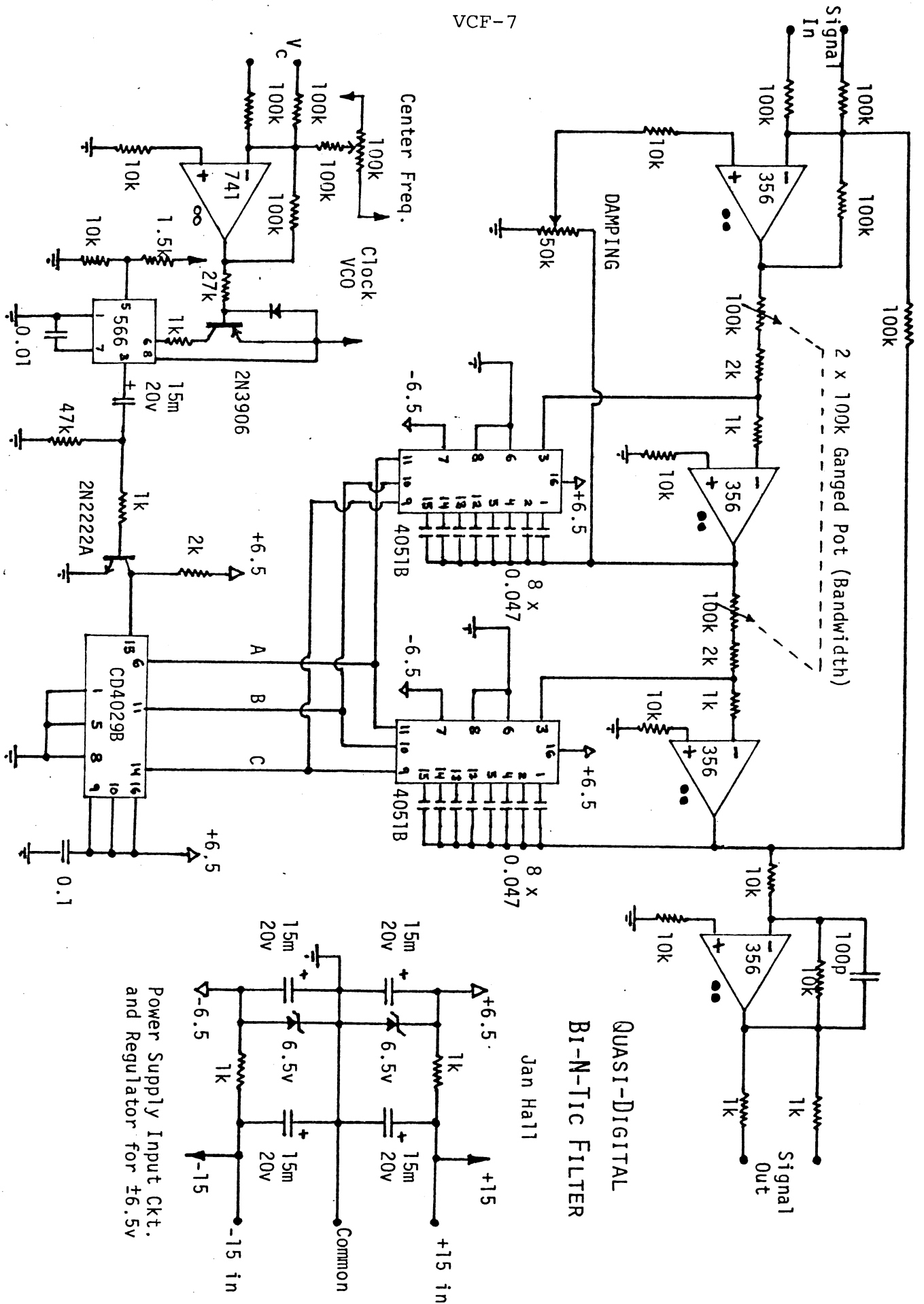


Fig. 1 CONTROL SUMMERS FOR ENS-76 VCF, OPTION 3, VARIABLE SLOPE FILTER



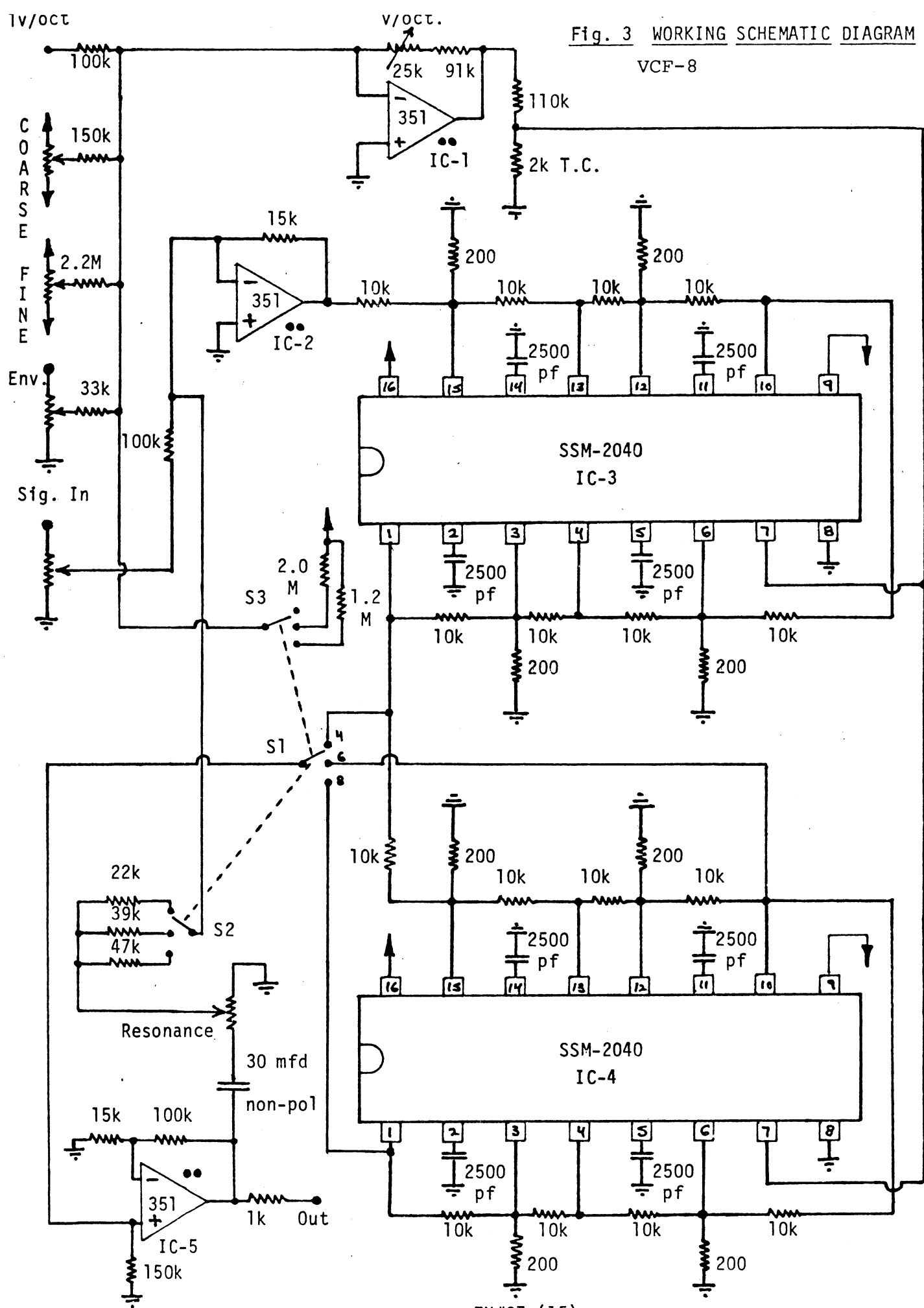




QUASI-DIGITAL  
BI-N-TIC FILTER

Jan Ha11

VCF-8





## TIMBRE MODULATORS

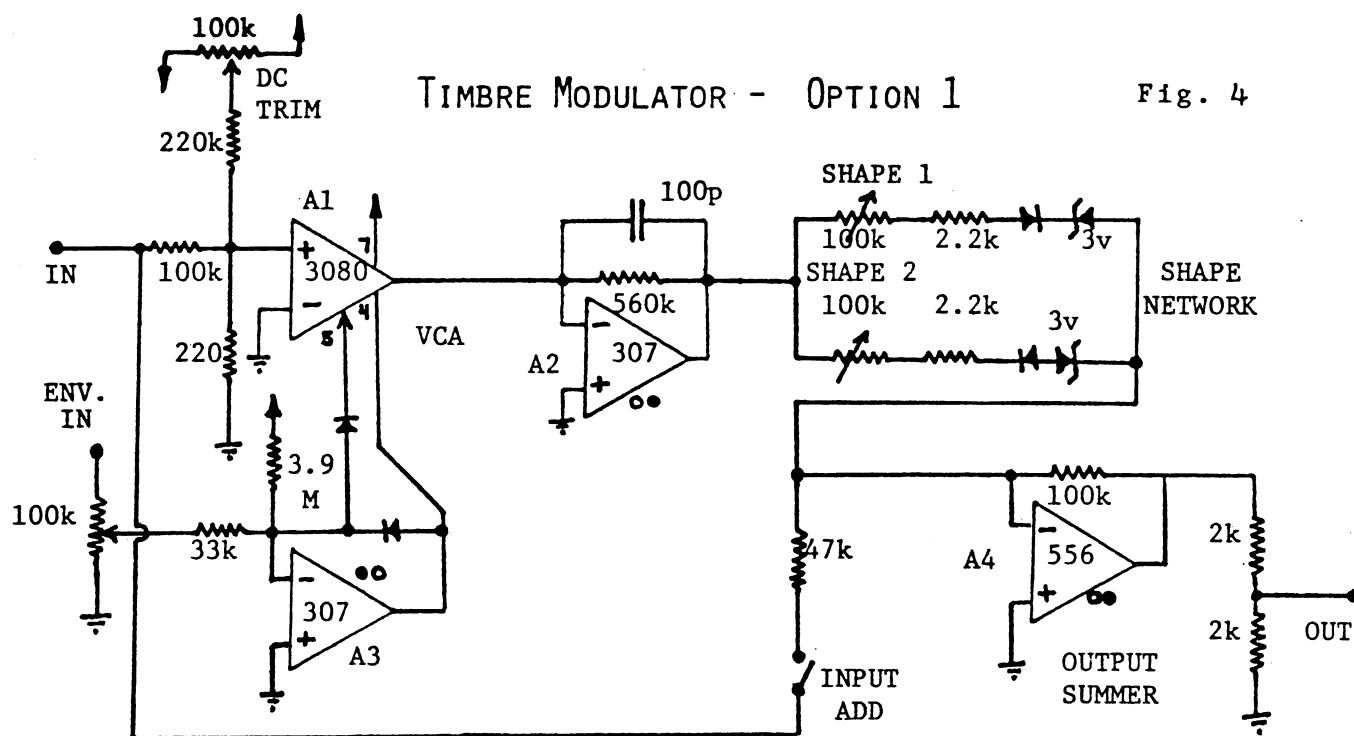
Timbre Modulators: These devices are circuits that perform in various ways the general function of enriching the tone color or causing the timbre to change. While the timbre modulation function is often performed with VCF's or with pulse width modulation, these newer timbre modulators offer additional effects, and often have an advantage of being waveform driven, and thus avoid expensive tracking exponential circuitry. Three modulators are offered in this collection.

| <u>OPTION</u> | <u>SOURCE</u> | <u>DESIGNER</u> | <u>APPROX COST</u> | <u>NOTES</u>  |
|---------------|---------------|-----------------|--------------------|---------------|
| TM-1          | EN#72         | Hutchins        | \$6                |               |
| TM-2          | EN#84         | Hutchins        | \$9                | Odd-Even Type |
| TM-3          | EN#72         | Fritz           | \$5                | Double Pulse  |



## TIMBRE MODULATOR - OPTION 1

Like a variable-slope filter, the "Timbre Modulator" is not your everyday type of synthesizer module. Also like a VCF, the timbre modulator is concerned with controlling the tone color of a signal. In fact, a VCF is a timbre modulator. The present device is a special type of timbre modulator - one which adds harmonics to a waveform of low harmonic content (such as a sine or triangle). If truth be known, we developed this circuit while attempting to do something else (which we will decline to mention to avoid looking unduly foolish). When we found that it imparted variable harmonic content to sines and triangles, it was obvious that we had arrived at a timbre modulator that was in many ways the counterpart of the VCF (which works best for sawtooth waves and pulses). The circuit is quite simple as can be seen by studying Fig. 4. It is basically a VCA which drives a zener diode input stage to an op-amp. The zener diodes prevent signals of less than about 3.5 volts from reaching the output of the module. Alternatively, the signal from the VCA can be added to the input for more complex waveforms. With the input added in, the most dramatic timber modulations are achieved, but the VCA action is blocked, so another VCA is needed somewhere in the circuit (and is of course, usually available).



TIMBRE MODULATOR - OPTION 1

Fig. 4

The VCA circuit of the timbre modulator is formed by A1, A2, and A3 in a manner similar to that described in EN#63. A couple of things have been added. The 3.9M resistor holds the VCA on slightly in the absence of an envelope. This assures that the CA3080 is not cut off when the envelope is low, as this condition would otherwise pin the output of A2. The capacitor in the feedback loop of A2 is just to prevent a slight oscillation that appeared on occasion, but otherwise has no function. The output of A2 drives the input signal (as controlled by the VCA) through the shaping network into the output summer A4. In order for a signal to get through this network, the amplitude must be in excess of about 3.5 volts. Thus we are center clipping the input waveform as can be seen from Fig. 5a which shows a sinewave input. Here we have assumed that the two shape controls are set to approximately the same value, otherwise the output would not be symmetrical. A case where the two controls are not set the same is shown by Fig. 5b.

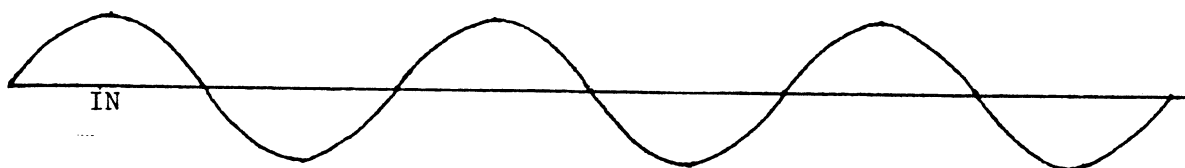


FIG. 5a SINE WAVE INPUT AND OUTPUT OF TIMBRE MODULATOR, ADD INPUT SWITCH OPEN

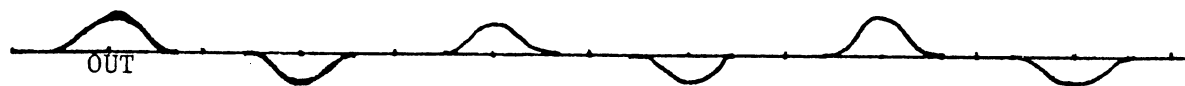
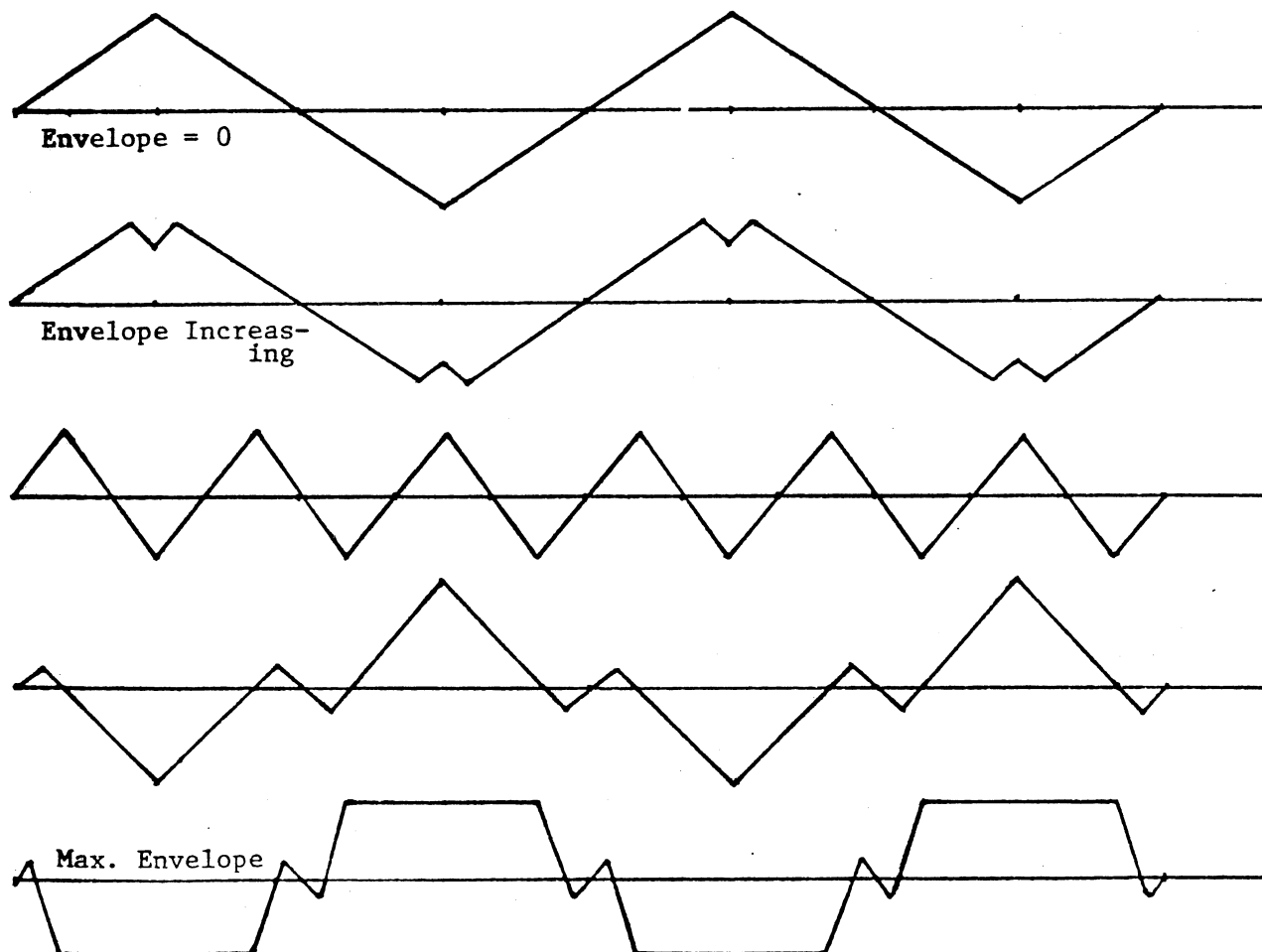


FIG. 5b SINE WAVE INPUT AND OUTPUT, ADD INPUT SWITCH OPEN, SHAPE CONTROLS UNEQUAL

By far, the most interesting effects with the timbre modulator are achieved with the INPUT ADD switch closed. With the switch closed, and the envelope at zero, the input waveform is passed to the output. Now, as the envelope rises, what is basically an inverted version of the input begins to appear at the output of A2. This would normally cancel the input if it were added to the output summer, but here it must first pass through the shape network. This leads to some very interesting results. An example of what one can expect is shown in Fig. 6 below.

FIG. 6 OUTPUT OF TIMBRE MODULATOR, ADD INPUT SWITCH CLOSED



When the envelope is low, the input triangle passes through unaltered. When the envelope rises, the first thing that will get through the shape network is the peak of the triangle. With the shape control pots set to about 10k each, this means that the peak of the triangle will start to cancel (and indeed cut deeper into) the peak of the input triangle, resulting in the waveform seen in the second line of Fig. 6. With a further rise of the envelope, we eventually arrive at the middle waveform of Fig. 6. While this is not a perfect waveform, it is very close to being a triple frequency version of the original triangle input. [This of course suggests a method of frequency tripling a triangle with a non-voltage-controlled version of this circuit.] As the envelope continues to rise, the output takes on the form shown by the fourth and fifth lines of Fig. 6. The fifth line is not unlike a square wave, and can be made to look more and more like a square wave by decreasing the resistance of the two SHAPE pots.

Thus, we have a device which starts with a waveform of low harmonic content, and greatly increases and alters this content as a control envelope rises. The effect is much like that achieved with a low-pass VCF and a waveform of high harmonic content. The harmonic evolutions are not the same as one gets with a filter however, so the timbre modulator should be a useful addition. It can also be used in parallel with a VCF (using the neglected sinewave output for example) and the two results can be mixed.

## AN ODD-HARMONIC TO EVEN-HARMONIC TIMBRE MODULATOR: -by Bernie Hutchins

TM-2

INTRODUCTION: Timbre modulators, of which the most popular type is the voltage-controlled filter (VCF) are very important modules in analog synthesis. Timbre modulators are devices which produce a changing tone color in response to a control (modulating) voltage. The timbre modulator described here is one which trades off between a waveform with all odd harmonics (square wave) and one with only even harmonics (full-wave rectified sine wave).

Before discussing the design of the timbre modulator, first we should review a bit the harmonic content of the waveforms we usually employ in analog synthesis. These are listed below:

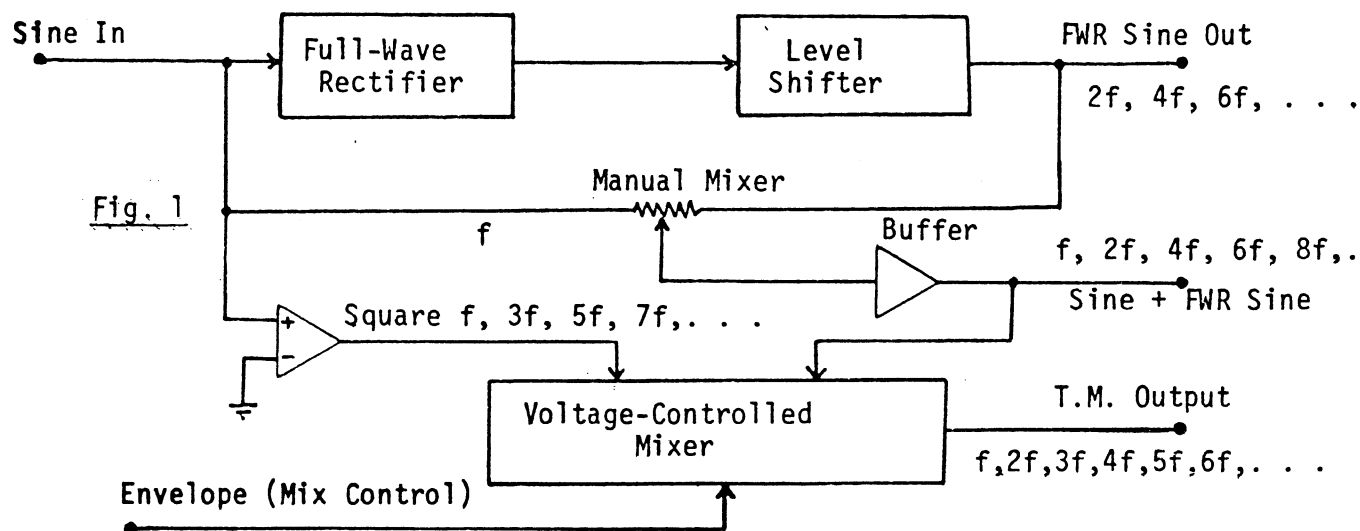
|                        |                  |
|------------------------|------------------|
| No Harmonics:          | Sine             |
| Odd Harmonics:         | Square, Triangle |
| Most or All Harmonics: | Sawtooth, Pulse  |

A curious fact emerges when we consider the "obvious" missing category - that of all even harmonics. This is because the fundamental (first harmonic) is odd, not even, and is therefore excluded! The remaining harmonics would be 2f, 4f, 6f, 8f, etc., which is exactly the same as a waveform of fundamental frequency 2f containing all harmonics. Thus, it does not make any sense to talk about a waveform with only even harmonics, unless we find some way to include in the fundamental somehow. We could look for a waveform with fundamental and all even harmonics (which we could obtain by Fourier synthesis if by no other way), but here it will be simpler to just use a waveform with only even harmonics and add in the fundamental by summing. It will be convenient to drive the module with a sine wave. This means that we can use the full-wave rectified sine wave as the waveform with only even harmonics, and the addition of the fundamental is a simple matter of summing the sine wave with the full-wave rectified version. Also, simple zero-crossing of the sine wave will give us a square wave, a waveform with only odd harmonics. Before going on to the block diagram of the timbre modulator, we will list below the harmonic component amplitudes of the three signals involved.

| <u>WAVE FORM</u> | <u>HARMONICS</u> |          |          |          |          |          |          |          |          |           |           |
|------------------|------------------|----------|----------|----------|----------|----------|----------|----------|----------|-----------|-----------|
|                  | <u>1</u>         | <u>2</u> | <u>3</u> | <u>4</u> | <u>5</u> | <u>6</u> | <u>7</u> | <u>8</u> | <u>9</u> | <u>10</u> | <u>11</u> |
| Sine             | 1.00             | 0        | 0        | 0        | 0        | 0        | 0        | 0        | 0        | 0         | 0         |
| Square           | 1.27             | 0        | 0.42     | 0        | 0.25     | 0        | 0.18     | 0        | 0.14     | 0         | 0.11      |
| FWR Sine         | 0                | 0.84     | 0        | 0.17     | 0        | 0.07     | 0        | 0.04     | 0        | 0.026     | 0         |

## BLOCK DIAGRAM OF THE TIMBRE MODULATOR

Fig. 1 shows a block diagram of the timbre modulator. Note that there are two inputs (the sine wave, and the envelope) and three outputs (FWR sine, mixed sine and FWR sine, and the final mixture with the square). Note that the lower outputs contain a portion of the upper ones.

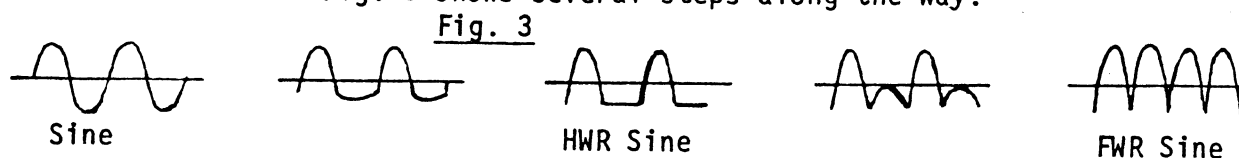


## CIRCUIT DIAGRAM OF TIMBRE MODULATOR

A full circuit diagram of the timbre modulator is shown in Fig. 2. The full-wave rectifier is realized with IC-1 and IC-2 while IC-3 serves as a level shifter and amplifier to produce a  $\pm 5$  volt signal as a Full Wave Rectified (FWR) sine. IC-4 forms the buffer. The voltage-controlled mixer is formed with two VCA circuits (IC-7 and IC-6, IC-12 and IC-9) which drive output stage IC-11, and are controlled by voltages adjusted by IC-10 and IC-8. All the circuitry is probably familiar from former ENS-76 circuitry or from the Musical Engineer's Handbook.

Note that we use the new LF351 op-amp exclusively here (except, of course, for the CA3080's). One surprise was the need for the capacitor C (10pf) to stabilize the first stage (IC-1) of the FWR. Apparently the instability of the stage without the capacitor C is due to the stray diode capacitance and low input impedance of the 351. All other stages were stable as expected, and the LF351 with its high speed proved adequate for the zero-cross operation (IC-5) to produce a square with sharp edges.

The first panel control to note is PC-1, which serves to adjust the form of the processed sine wave which will be fed into one channel of the voltage-controlled mixer. This signal is also brought out through R10 in the event that the manually controlled timbre may be useful. Note that you can adjust PC-1 to give the same FWR sine output that is available through R8, so if you wish, you can save one panel jack by eliminating the FWR sine output through R8. It is also interesting to note that in addition to the FWR sine and the sine itself, the pot PC-1 allows the user to obtain a continuous mixture, including the Half-Wave Rectified (HWR) sine at the midpoint of the mix. Fig. 3 shows several steps along the way.



Two other front panel controls are available, PC-2 which sets an initial timbre and PC-3 which adjusts the level of an external envelope to vary the timbre. In normal operation, the voltage at the output of IC-8 (the control for the upper VCA, IC-6 and IC-7) varies from 0 to +5 exactly as the sum of the voltages on the wipers

of PC-2 and PC-3. At the same time, the voltage at the output of IC-10 (the control for the lower VCA, IC-12 and IC-9) varies from +5 to 0. Thus the sum of the two control voltages to the two VCA's is always +5 so the gain of the two VCA's sum to one. When the control voltage sum is positive approaching +5, we have mostly square wave coming through, and when the control voltage sum approaches 0, we have mostly the processed sine. For example, if PC-1 is in the sine position, and the control voltage sum varies from 0 to +5, the waveform at the Timbre Modulator Out (output through R22) will change from a sine to a square. Fig. 4 shows some steps along the way during this transition.

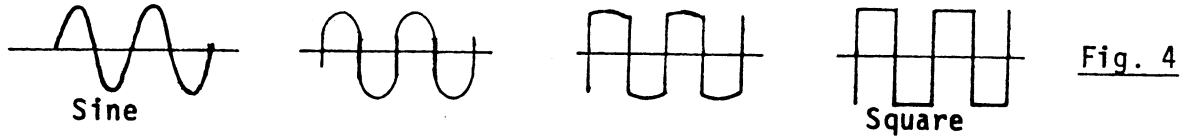


Fig. 4

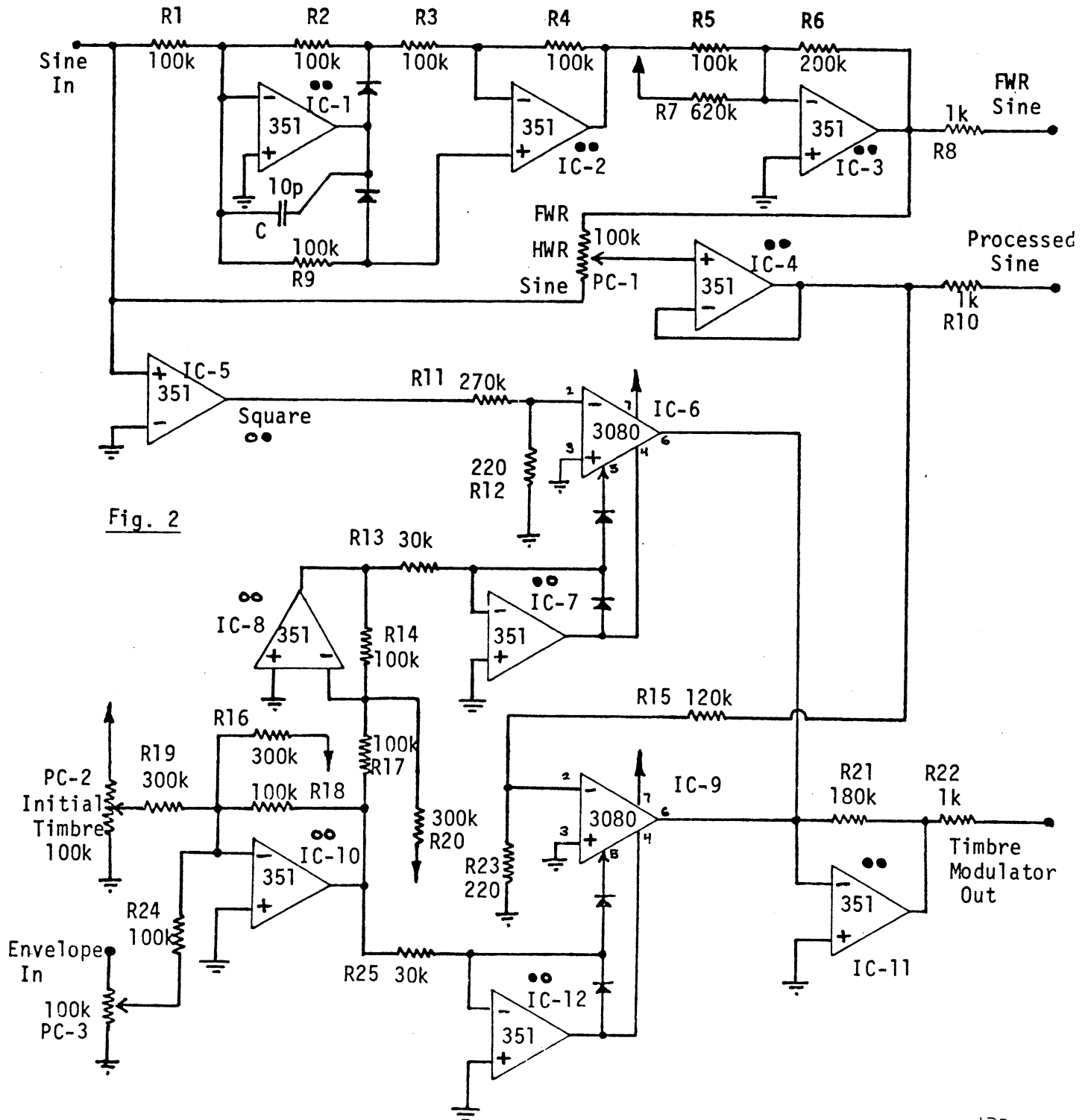
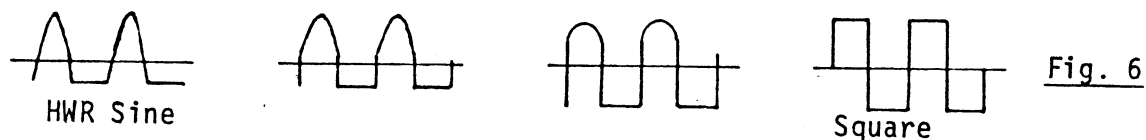
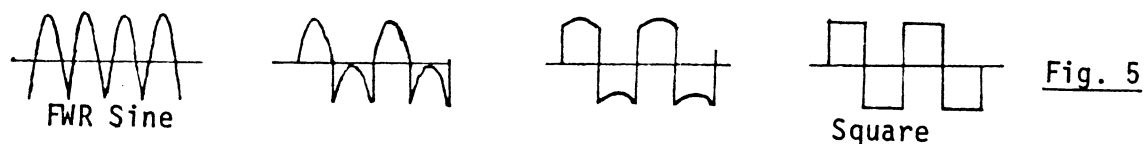


Fig. 2

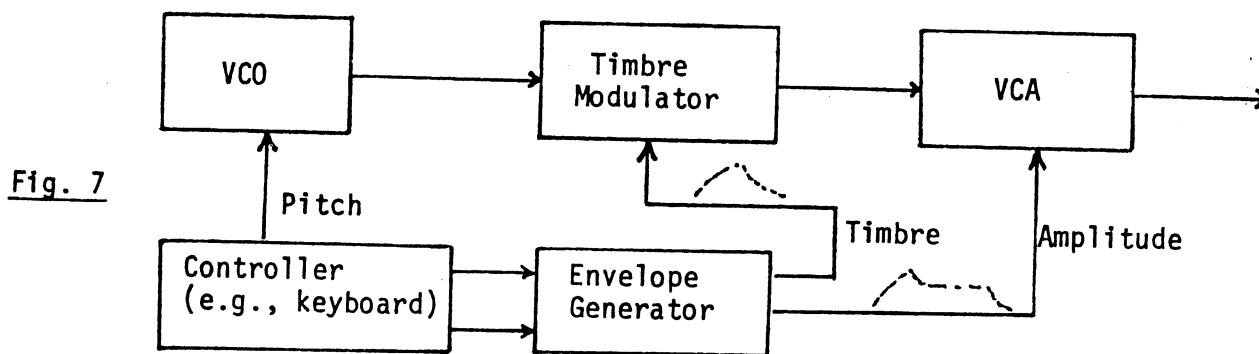
We can also consider how the waveform is transformed when the processed sine is a FWR sine (Fig. 5) and a HWR sine (Fig. 6).



While it is interesting to be able to produce different types of waveforms (in particular, as low-frequency control waveforms), when it comes to timbre, we are more interested in the changing timbre that results from a changing harmonic structure than we are in waveshape. As discussed in the introduction, it is the transition from even to odd harmonics and back again that is of interest. At the same time, when we go over to talking about musical timbre, it becomes more difficult to describe the results in other than purely subjective terms. Perhaps it will mean something to some readers if we say that the present timbre modulator has about the same total effect as a 12db/octave filter, but the transition is somewhat brighter, probably due to the strong octave content of the FWR sine. Of course by adjusting the controls, a wide range of effects are achieved, including some rather subtle ones.

### USING THE TIMBRE MODULATOR

The timbre modulator is typically employed in much the same manner as any other timbre modulator such as a VCF. An example application is shown in Fig. 7 where the timbre modulator assumes the VCF position.



Several important features of timbre modulators of this type can be ascertained from Fig. 7. One important feature is the absence of a pitch control voltage path from the controller to the timbre modulator. Typically, the VCF needs this path and must be able to track the VCO. The added circuitry required is well known for its expense and need for fine tuning. Another feature of the timbre modulator is that while we designed it around a sine wave input, we could use triangle, sawtooth, or other waveforms that do not have all sharp transitions to drive it. It would still produce timbre modulation. Note that since we can start with a sine wave, the timbre modulation in this case is not subtractive synthesis as with the VCF. We can add harmonics that are not present in the input. This is of course due to the non-linear element in the timbre modulator (in this case, the FWR). A VCF on the other hand is a linear processor, so a sine wave in can only give a sine wave out (albeit attenuated and phase shifted). This last parenthetical remark leads us to a final point about this type of timbre modulator: it does not result in a phase shift as a filter does. Whether this is a good point or a bad point depends on what you are using it for. Of course, there is no reason why this type of timbre modulator can't be used with VCF's as well.



Ian Fritz has submitted three waveshaping circuits which are somewhat different from those generally used, and thus should be of interest to persons using waveshape as a principal method of timbre control. The first circuit is shown in Fig. 2 and is a pulse-width modulator for a "double pulse" rather than the usual single pulse. The second circuit (Fig. 3) is a shaper circuit which squares up a sine wave without actually turning it into a square wave (it rounds off the corners). The third circuit (Fig. 4) can be used to either chop off the bottom half of a sine wave, or to cut back on the amplitude of the lower half, according to the setting of the control pot. This results in the generation of even harmonics which would be absent from many shapers (such as the one in Fig. 3 for example).

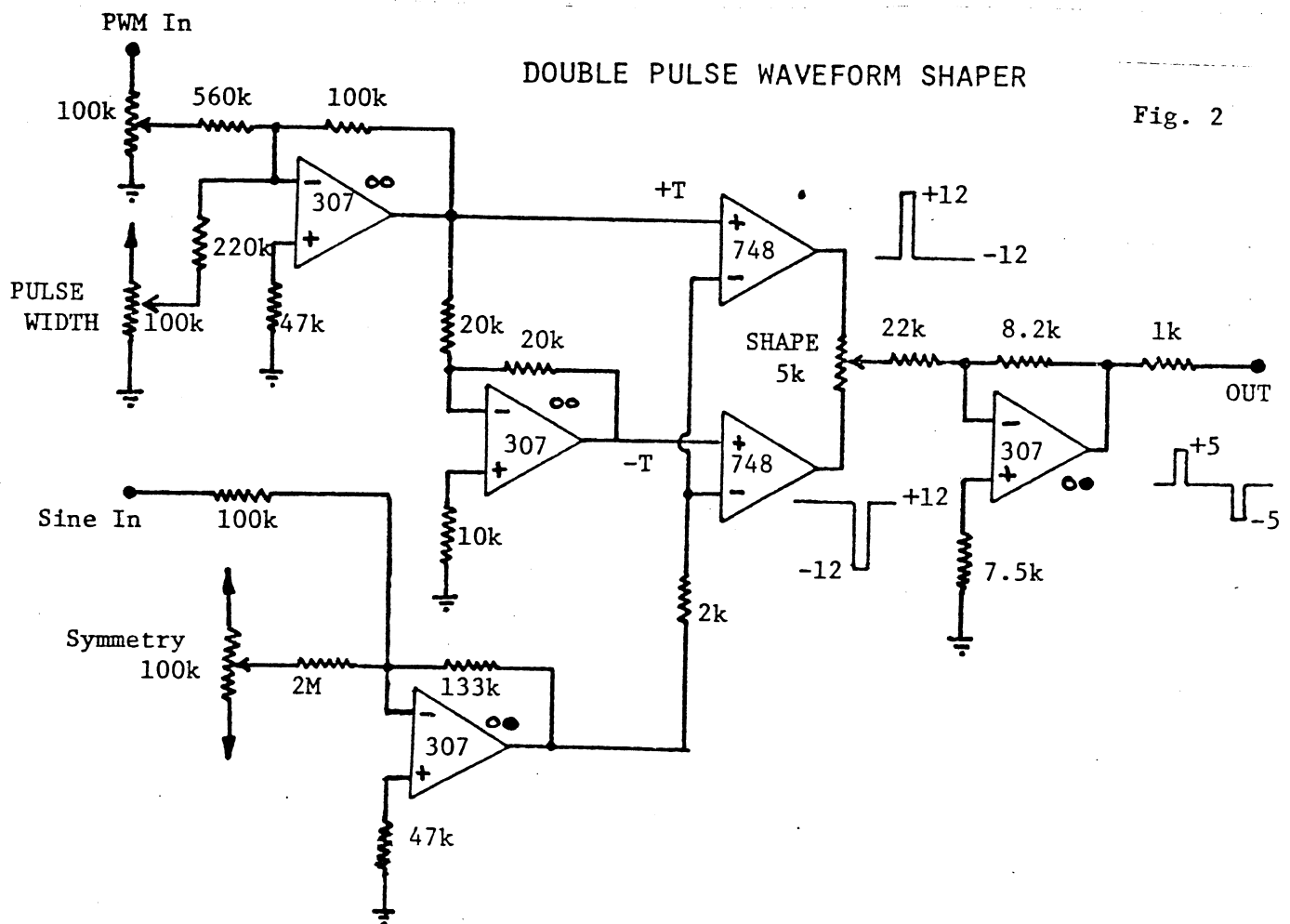


Fig. 2

EN#72 (21)



## ENVELOPE GENERATORS

THE ENVELOPE GENERATOR OPTIONS: We start the envelope generator options with two options from the ENS-76 series. The first option is a gate and triggered standard ADSR generator, and can be built from four IC chips. The second option is the same as the first except a delay section is added. We also show some accessory circuits that can be used to obtain an instantaneous adjustment of the sustain level in these generators. The third and fourth options given here are older circuits which may well still be of interest. Option 3 is an AD type of envelope generator that provides a "blip" type of envelope in response to a trigger. Option 4 is our older ADSR design based on a slightly different principle - the addition of an AD and an AR type of envelope. This makes possible some slightly different envelope shapes, and a couple of other options such as triggering of the AD section and simultaneous outputs of three different envelope shapes. Option 5 is a very simple ADSR since the main part of the circuitry is on the SSM-2050 IC chip. In addition, it is totally voltage-controlled, and is the best option if voltage-control is needed. Option 6 is an unusual design - a gate driven 2nd order filter, which provides what may be some more "natural" envelope shapes. Option 7 includes a "muting" feature but is otherwise an AD + AR type.

| <u>Preferred<br/>Circuit<br/>Option</u> | <u>Original<br/>Source</u> | <u>Original<br/>Designer</u> | <u>Estimated<br/>Parts<br/>Cost</u> | <u>Notes</u>     |
|-----------------------------------------|----------------------------|------------------------------|-------------------------------------|------------------|
| EG-1                                    | EN#66                      | Hutchins*                    | \$5                                 | ENS-76 Option 1  |
| EG-2                                    | EN#66                      | Hutchins*                    | \$7                                 | ENS-76 Option 2  |
| EG-3                                    | MEH 5e (4)                 | Hutchins                     | \$3                                 | AD generator     |
| EG-4                                    | EN#45                      | Hutchins                     | \$6                                 | ENS-74 Option 1  |
| EG-5                                    | EN#87                      | Rossum                       | \$9                                 | Uses SSM-2050 IC |
| EG-6                                    | EN#86                      | Hutchins                     | \$9                                 | 2nd Order Filter |
| EG-7                                    | EN#92                      | Fritz                        | \$8                                 | with Muting      |

\*includes a logic scheme first given us by Dave Rossum



## THE ENS-76 HOME-BUILT SYNTHESIZER SYSTEM - PART 2: -by Bernie Hutchins

In this installment, we will be looking at envelope generator designs. On the one hand, envelope generator circuits tend to be fairly routine. On the other hand, they also tend to be tedious to analyze - there is generally one overall system that has a multitude of possible states and sequences. For these reasons, we want to start right out here with a short description of what is to follow. This will serve to point out the new features of these circuits in a way that does not involve a detailed analysis.

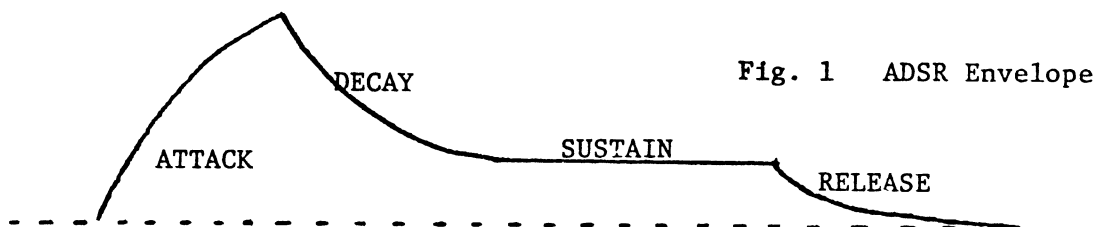
The first thing we will do is present the first circuit (Option 1) which is actually the circuit from EN#50 with a small amount of reworking. The new feature is a switching arrangement that makes it easier to use as an AD type generator.

Then we will add on a delay unit to Option 1 to form Option 2. This is basically the delay circuit from EN#51 with a few rearrangements of gates and buffers. We have also cleaned up the switching sequence to prevent glitches that may be a problem in some cases.

The third thing we do is to describe an option that can be added to either of the first two circuits (and probably to others) that gives an instant response feature to the sustain control. Consider that when you set the sustain level you are usually varying a sustain level voltage and that the output of the envelope generator varies with the time constant of the decay circuit. A small bother, but here is a way of avoiding it if you want to. With this option, you get instantaneous response.

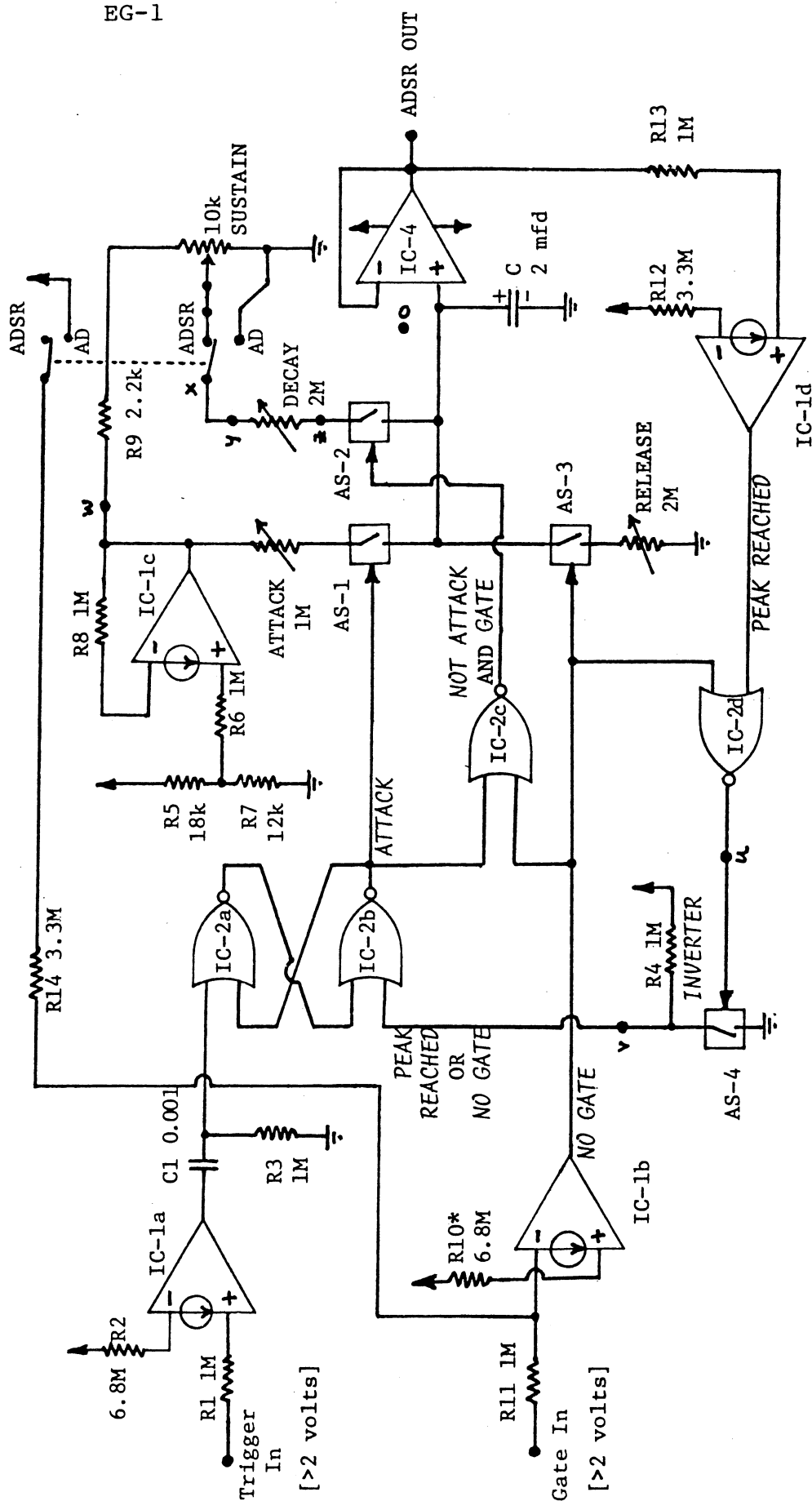
### ENS-76 ENVELOPE GENERATOR - OPTION ONE

Option 1 in the envelope generator series is shown in Fig. 2. It is essentially the "four-chip" ADSR (Attack-Decay-Sustain-Release) generator that was first described in EN#50. The ADSR envelope waveform is indicated in Fig. 1 below for reader's who may be unfamiliar with this envelope which has become fairly standard.

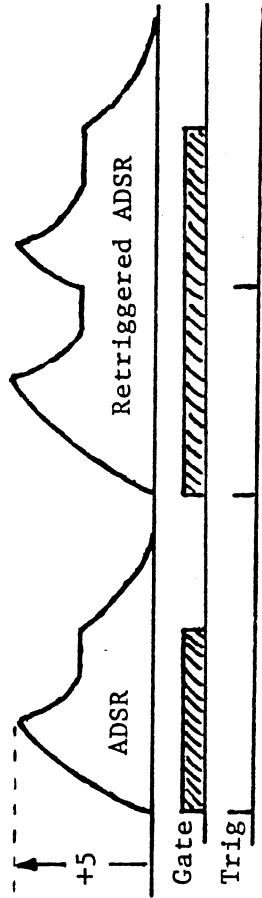


The basic theory of the ADSR envelope generator can be found in several places including Dave Rossum's description in EN#22 and in Chapter 5e of the Musical Engineer's Handbook. For this reason we will not go into a lot of description here. We will look at two models of the ADSR which will be useful when we get to voltage-controlled designs. First we should observe that a waveform of the type shown in Fig. 1 can be produced by a simple RC charging circuit with certain sequencing circuits that determine the charging (or discharging) voltages and time constants. This idea is illustrated by the diagram in Fig. 3.

EG-1



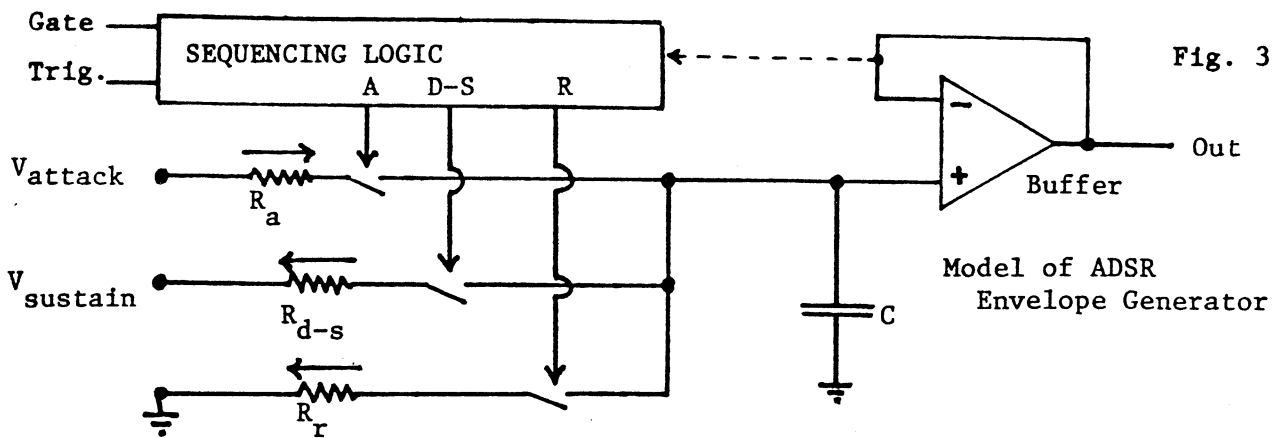
OUTPUT ENVELOPES AND TIMING



\*This resistor is mislabeled in some copies of EN#50 and in MEH Chapter 5e (7) as 6.8 rather than 6.8M. Please make corrections.

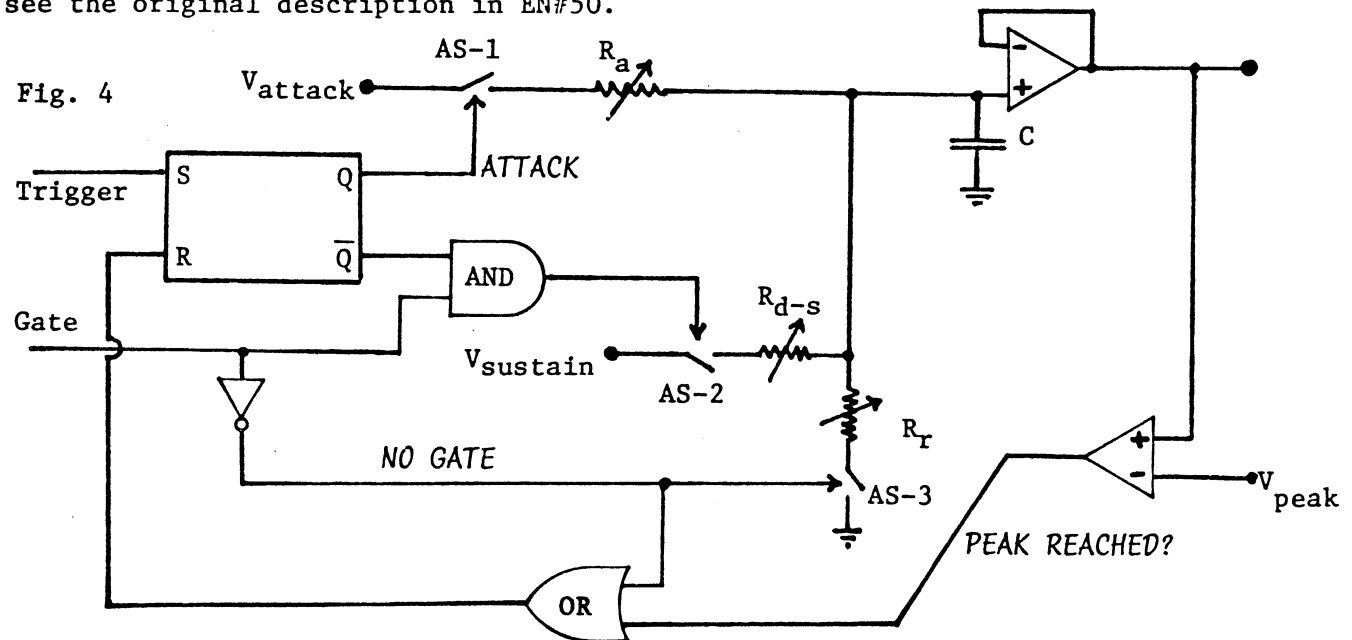
Integrated Circuits:

- IC-1a,b,c,d LM3900
  - IC-2a,b,c,d 74C02 CMOS
  - IC-3 (AS-1, AS-2, AS-3, AS-4) 4016 CMOS
  - IC-4 307 (or 741)
- Note: All IC's except IC-4 powered between +15 and ground. IC-4 between +15.



In the above model, we show three principal sections. First there is a capacitor with an attached buffer. Secondly, there is some sequencing logic which controls appropriate analog switches. Finally, there are certain reference voltages ( $V_{attack}$ ,  $V_{sustain}$ , ground) which are available through corresponding resistances.

One form of the sequencing logic is shown by the setup of Fig. 4. This is the usual combinational logic for an ADSR system that is controlled by gate and trigger commands from a controlling device (e.g., a keyboard). The reader can easily observe how the logic is used. For example, the decay-sustain mode is determined when a GATE is present AND when the attack mode is not in effect. A full description of this logic scheme is found in the references given above. The circuit in Fig. 2 is an implementation of the logic of Fig. 4. For a full description of the operation of the circuit in Fig. 2, see the original description in EN#50.



The one difference between the circuit in Fig. 2, and the earlier EN#50 circuit is that this circuit has a switch that is useful for using the AD mode only. It is well known that an ADSR envelope generator can be used as an AR (Attack-Release) generator by just setting the sustain level to the maximum (peak) value. Also, the ADSR can be used as an AD (Attack-Decay) generator (no sustain; thus independent of gate) by setting the sustain level to zero. However, note that while an ADSR used in this manner produces an AD envelope that is independent of the gate as far as its timing is concerned, there must be a gate present for the AD to occur. With the addition of the ADSR-AD switch shown in Fig. 2, two things occur. First, enough current is dumped into the gate comparator (IC-1b) through R14 to keep this comparator low (actually the no-gate condition, but the comparator looks for no-gate rather than gate). In other words, it makes the

envelope generator think it always has a gate. The other half of the switch can be seen to set the sustain voltage to zero whenever the AD mode is employed. The way this switch might be useful is illustrated by the following example. Suppose we are using a delay unit giving 10 seconds delay. If the switch is in the ADSR mode with  $S=0$  volts, we make AD type envelopes. Now suppose we press a key down and hold it. Ten seconds later a note sounds under the AD envelope. If however we release the key after three seconds, no note will sound. With the switch in the AD position however, it is not necessary to hold a key down. We can tap a key, walk across the room, and the AD envelope will still occur after ten seconds. There are other examples that could be given where this switch can be very useful.

## ENS-76 ENVELOPE GENERATOR - OPTION TWO

In Option 2 (Fig 5, next page) we add a delay unit on to the Option 1 circuit. To do this we use the basic circuit from EN#51. For a full description of the circuit and timing diagrams, see the EN#51 presentation. The original circuit was used directly connected to digital logic levels from a keyboard interface. In this reworked version we have removed the buffers from the input of the ADSR and put them on the inputs of the delay unit instead (which makes more sense in the general case). Thus, the LM3900 CDA's that served in Option 1 are moved to the delay unit while one of the inverters (part of the 74C02 quad NOR gate), IC-5d moves down to give the "NO GATE" signal to the ADSR. In this way, option 2 requires only 6 IC's total. The delay unit is operated on +15 volts so the 74C02 CMOS chip is used rather than the optional 7402 TTL chip in the earlier design. However, since the inputs are buffered and respond to any signals that exceed about +2 volts, this circuit is easily driven by TTL levels or by CMOS. The maximum delay with this circuit as drawn is about 4 seconds. This can easily be increased by increasing the DELAY TIME pot to about 10M, or by increasing C6.

In addition to the rearrangement of buffers and gates, we have added two devices intended to reduce erratic triggering. The first of these is the 0.1 mfd capacitor C2 which bypasses the power supply on IC-6. The inclusion of such capacitors is more or less standard practice when type 555 timers are used, and this one should not be omitted. The second thing is the RC combination R16-C3 which gives about a 1 ms delay to the undelayed gate signal. This assures that no "glitch" will occur if it takes a little time for the delaying process to set up (through IC-5b, IC-5c, and IC-6). A glitch coming through could trigger some envelope generators, including this one in the AD mode. [Actually we tried it without the R16-C3 network and there was no problem, but it is probably safest to leave it in.]

## ADDING "INSTANT SUSTAIN LEVEL ADJUSTMENT"

When we first started making ADSR envelope generators, we used a method in which an AD envelope was added to an AR type. This worked well, but had the drawback that it was necessary to use a dual pot so that both attack sections were adjusted the same. Such an ADSR generator can be found in the ENS-74 system, EN#45 (15). The ADSR circuits as in Options 1 and 2 of this issue are probably easier to build. In going over to these new designs, there were two features of the earlier design that we missed. The first was the fact that the AD section of the new circuits was not independent of the existence of a gate signal. This has been corrected by the addition of the ADSR-AD switch. The second thing was that the sustain level of the new circuit had a time delayed response as the control was adjusted. This was not present in the ENS-74 design, and it is our purpose here to show how the time delayed response can be removed from the newest designs.

This time delayed response can be understood by studying the circuit in Fig. 6.

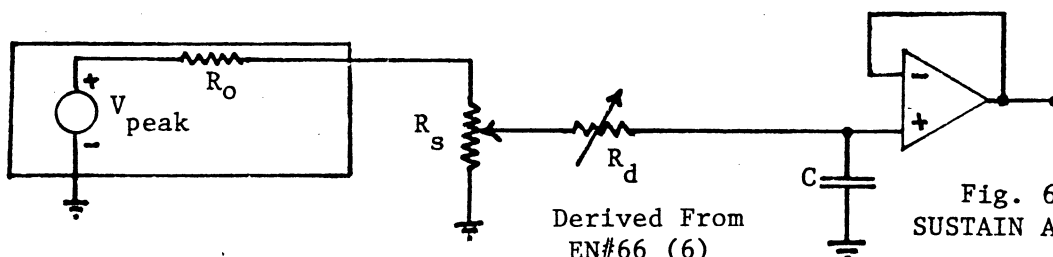


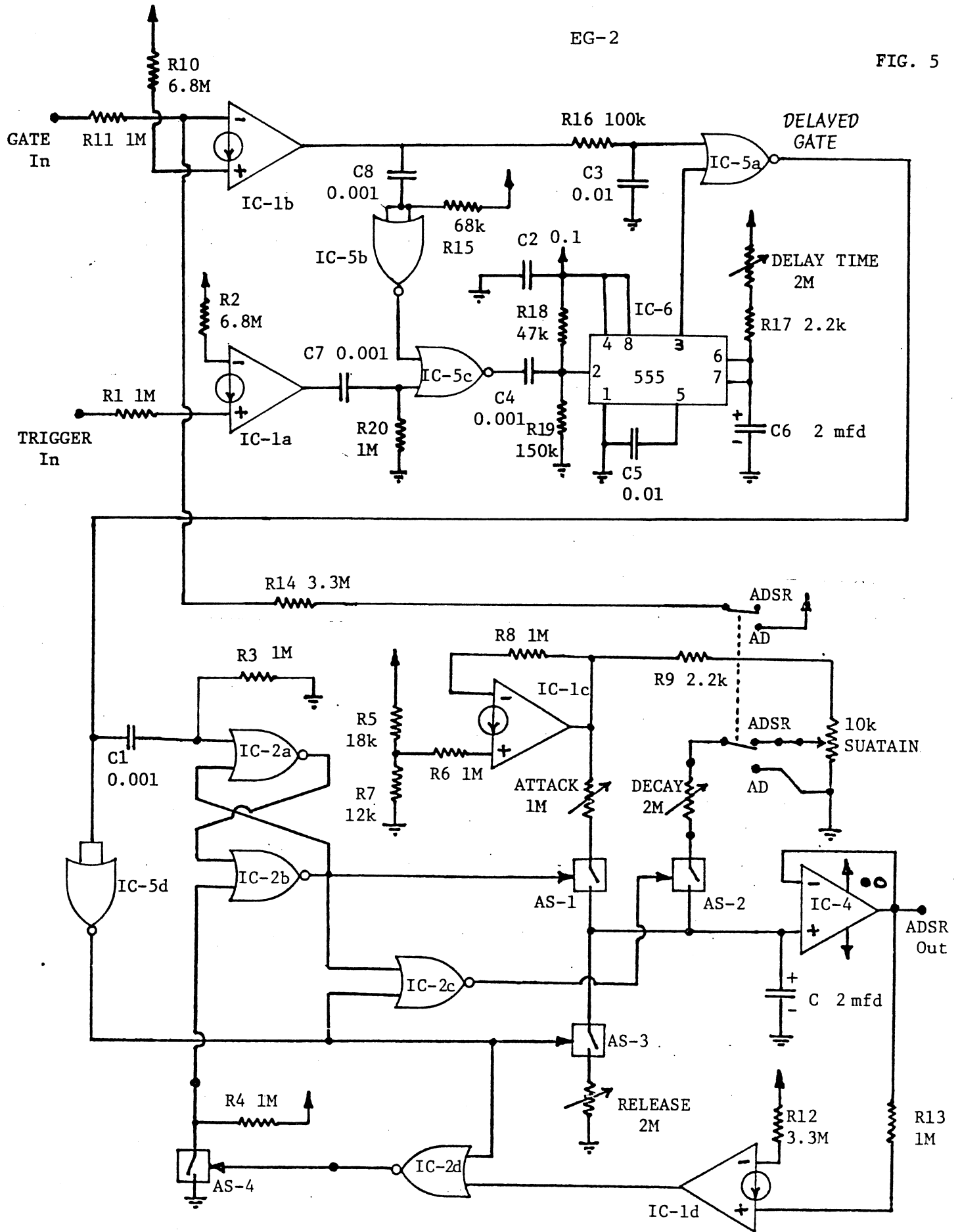
Fig. 6  
SUSTAIN ADJUSTMENT



# ENS-76 ENVELOPE GENERATOR OPTION TWO

EG-2

FIG. 5



EN#66 (7)

We can make the following assumptions about the above model. First, the output impedance of the source of the voltage  $V_{\text{peak}}$  (or  $V_{\text{attack}}$ ) is generally fairly small (about 200 ohms in the case of the LM3900 source in Fig. 2). Secondly, we know that the time constant of the  $R_s$  pot and the capacitor  $C$  is approximately  $(R_s/2) \cdot C$  or less, and this is most likely only a few milliseconds (10 ms in the case of Fig. 2). Thus, the only noticeable time delay is likely to result from a high resistance setting of the  $R_d$  pot. In the circuit of Fig. 2 for example,  $R_d$  may be 2M and  $C$  is 2 mfd, so a 4 second time constant results. What does this mean? Well, suppose you want to set the sustain level of the ADSR generator. You press down a key on the keyboard and wait for the envelope generator to cycle through attack and decay. When the sustain level settles down, it may well not be the one you want. You then turn the  $R_s$  pot up or down. Instead of responding instantaneously, there is a delay as the voltage on the capacitor can change only as current moves through  $R_d$ . In fact, it takes one time constant for the output to respond by about 2/3 of the difference. It may thus take several time constants for the new sustain level to be established with enough accuracy. The process of course does converge, and if necessary the  $R_d$  pot could be set to zero during the adjustment, but most users would agree that everything else being equal, it is desirable to have the sustain control have no time delayed response at all.

Thus we will discuss the necessary circuitry here. The builder must decide if the additional circuitry can be justified. Two points against should be brought out. First, if the user is using mainly short decay times (short notes), then the time constants are small and probably little problem. Secondly, the user often has to evaluate the sustain setting "on the fly" anyway, and does not just stall the generator in sustain and adjust the control.

Assuming you do want this feature, there are several ways to go about it. Obviously you want to short out  $R_d$  during the time the control is being adjusted. One way would be to use a metal knob on a plastic shaft for the  $R_s$  pot, and rig up touch control circuitry. Thus, when the knob was touched, an analog switch could be closed to short out  $R_d$ . Another approach, which we shall use here, is to use a differentiator on the pot wiper voltage to determine when the pot is being adjusted, and then use this information to close an analog switch. A circuit which we have tested and found successful is shown in Fig. 7. In this scheme, IC-1 buffers the pot voltage in the event that the impedance is too high (don't use it if you don't need it). IC-2 forms a high gain differentiator and IC-3 forms a full-wave rectifier (non-precision). When the pot is adjusted (rotated), the change in wiper voltage forces the output of the differentiator into (+ or -) saturation. This is rectified by IC-3 and then turns on the analog switch, shorting out  $R_d$ . If the pot is adjusted very slowly, there is not enough output from the differentiator to trigger the analog switch, but by the same token the change is small and the proper sustain level converges rapidly. Naturally, this works best when the pot is adjusted in small rapid motions.

When this circuit was first breadboarded, it worked just great. A second setup did not work so well, and we had to increase the gain of the differentiator to get it to work satisfactorily. After a little checking, it was found that the difference was due to the fact that the sustain pot in the breadboard was wire-wound while the one in the second setup was carbon. It seems that the wire-wound pot "bounces" quite a bit and one need not rely on just the  $dv/dt$  due to the pot rotation. In fact, the wire-wound pot seems to bounce so much that you get saturated noise at the output of the differentiator that is the same with either direction of rotation. In fact, there was so much output that we could just use the half-wave rectifier shown in the lower portion of Fig. 7. Extending this idea even further, it is possible to add instant sustain level adjust to either of the options given earlier in this issue without using any more IC's. To do this, we free up one analog switch and one of the LM3900 CDA's. This is done by using a transistor inverter in place of AS-4 and a 6 volt Zener diode in place of IC-1c. The circuit is shown in Fig. 8 below:

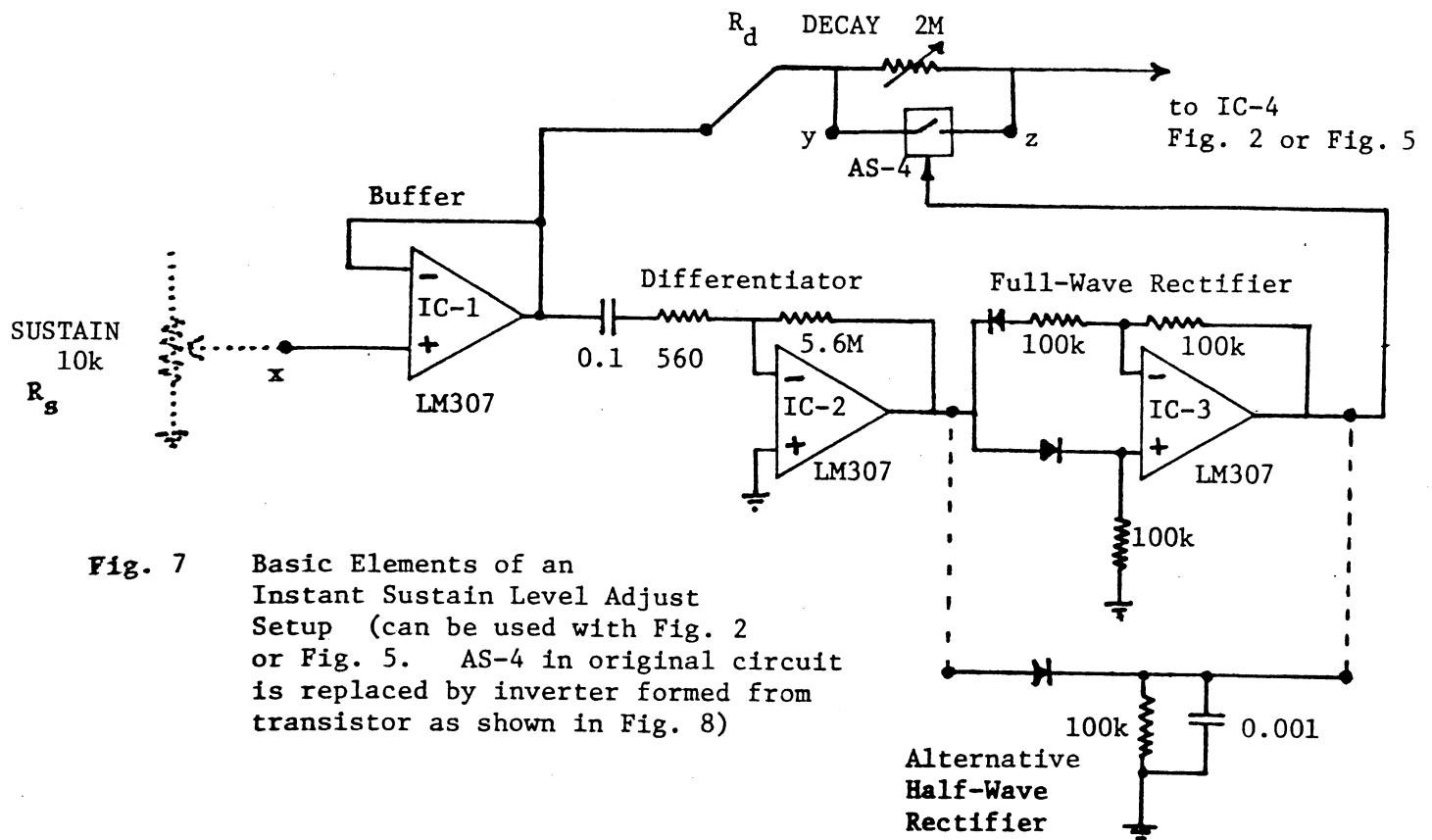


Fig. 7 Basic Elements of an Instant Sustain Level Adjust Setup (can be used with Fig. 2 or Fig. 5. AS-4 in original circuit is replaced by inverter formed from transistor as shown in Fig. 8)

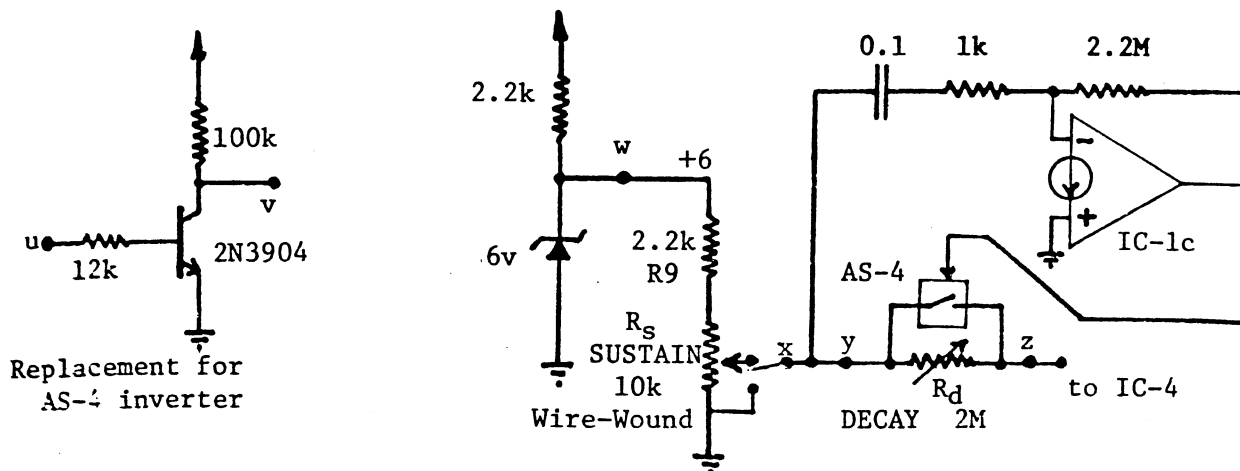
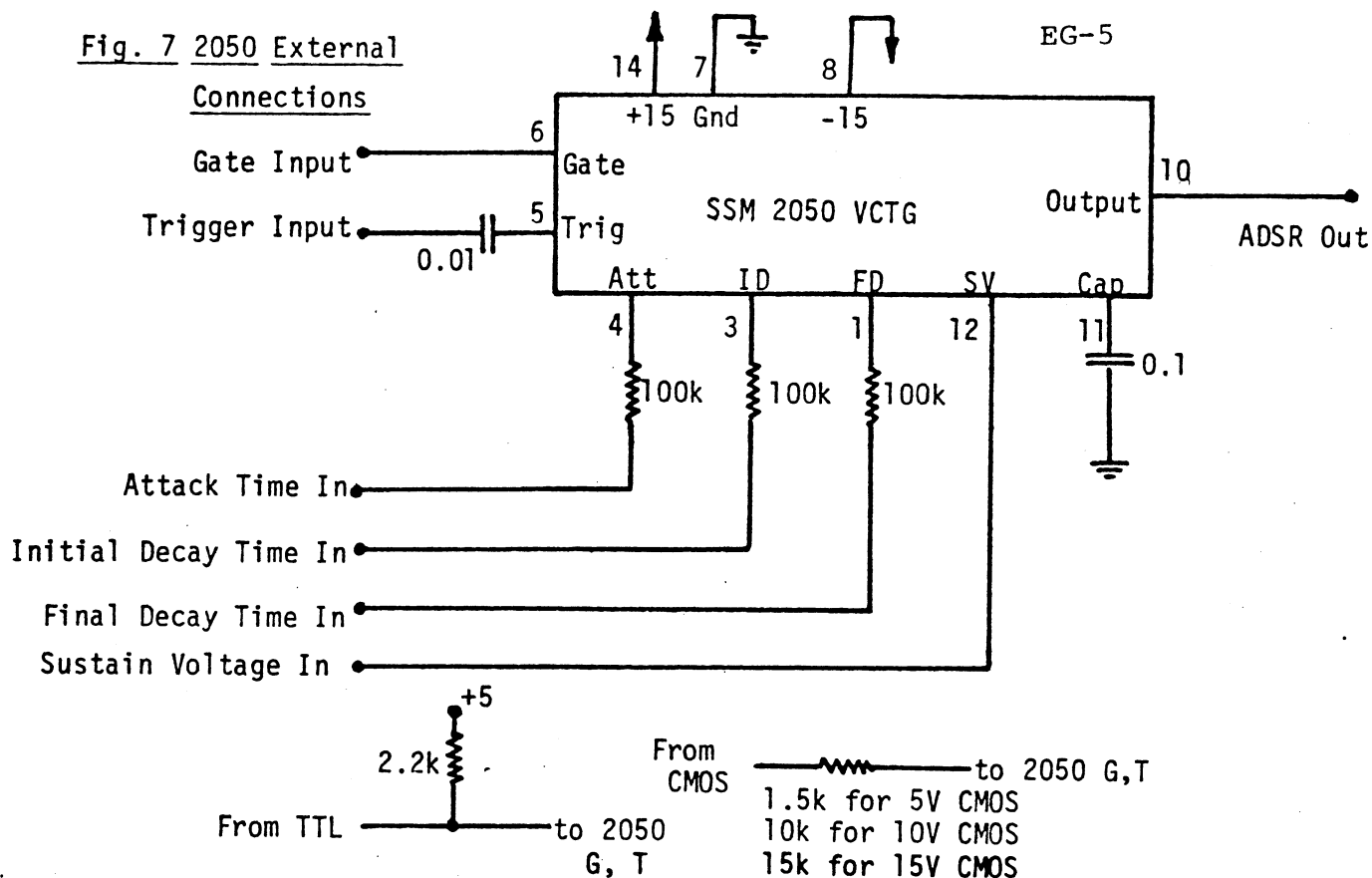


Fig. 8. Alterations to Fig. 2 or Fig. 5 for Instant Sustain Adjust Feature.

With the arrangement of Fig. 8, a system is achieved that works very well and does not require additional IC's. The output of IC-1c is "hash" that appears whenever the pot wiper is moved. This "hash" varies between +15 and ground, but is high enough for enough of the time to turn on AS-4 part time and thus short  $R_d$ . This is so simple to include, it should be made a part of either option as long as there is no problem getting the wire-wound pot required for  $R_s$ .



Fig. 7 2050 External Connections



The circuit connections for the 2050 (Fig. 7) are simple: a connection to each time input through 100k gives 2V/decade sensitivity. The sustain voltage is a voltage input. The trigger should be capacitively coupled to a source of a rising edge, and the gate directly to a DC level. Fig. 7 also shows interfaces to various logic families.

As the 2050 initial rates vary over a 1:4 range, and the time input sensitivities vary over a 60% range, some selection and trimming may be necessary for matching units, as may be required in a polyphonic system. The input sensitivities will be matched within the chip for the three time inputs, and can be measured by the resistance measured at low current from pin 1 to pin 7. Matching chips by this resistance will result in units with matched input sensitivities. The initial rates could be matched by selecting capacitors, but a single trimmer diividing  $\pm 15$  connected to all three time inputs via three 2.2M resistors will trim the initial rate for the chip; again the initial rates for all three inputs are matched within the chip.

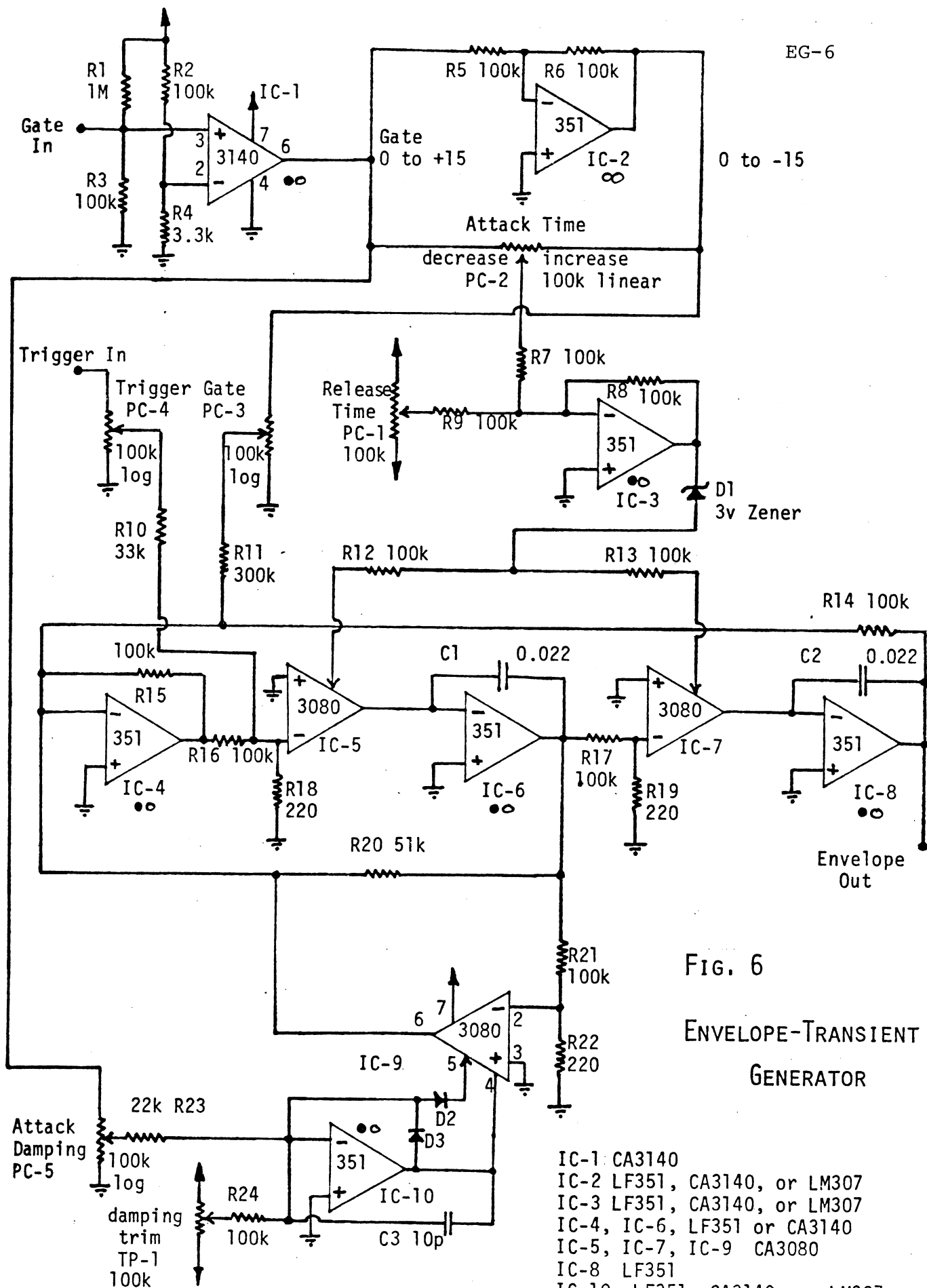
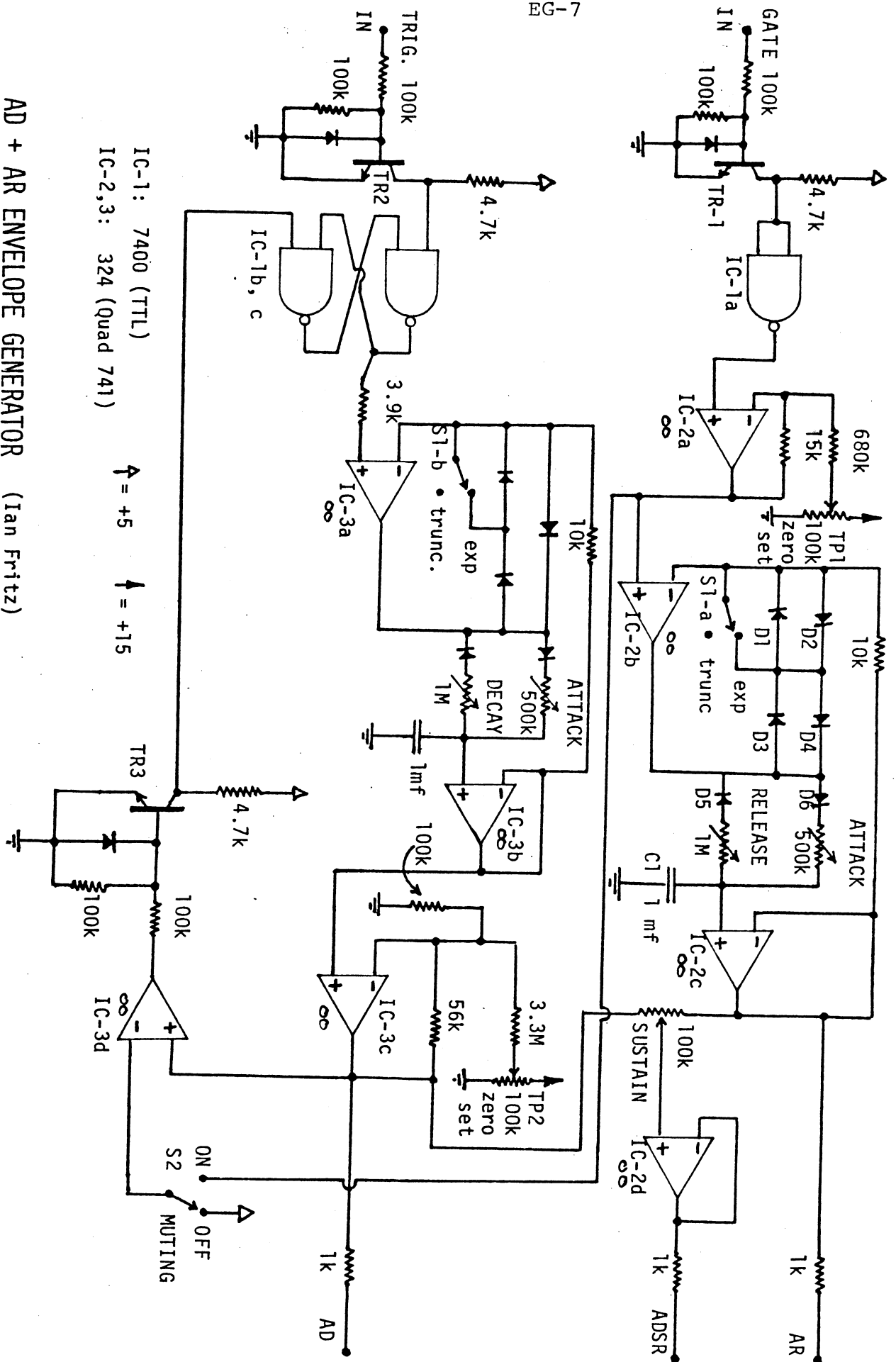


FIG. 6

# ENVELOPE-TRANSIENT GENERATOR

IC-1 CA3140  
 IC-2 LF351, CA3140, or LM307  
 IC-3 LF351, CA3140, or LM307  
 IC-4, IC-6, LF351 or CA3140  
 IC-5, IC-7, IC-9 CA3080  
 IC-8 LF351  
 IC-10 LF351, CA3140, or LM307  
 D2, D3 1N914, 1N4148, etc.

## AD + AR ENVELOPE GENERATOR (Ian Fritz)







## BALANCED MODULATORS

THE BALANCED MODULATORS (RING MODULATORS): Four circuits for balanced modulators are given in this collection. All four are full four-quadrant multipliers, while the fourth has a VCA option as well. The first option uses two CA3080's working back-to-back as two-quadrant multipliers to produce the full four-quadrant multiplication. The second option uses the older but still useful type 595 four-quadrant multiplier. The third option is a top of the line design using the excellent AD533 multiplier from Analog Devices, Inc. The fourth option uses the LM13600 OTA device in a balanced modulator/VCA option and is a good choice for a beginning system.

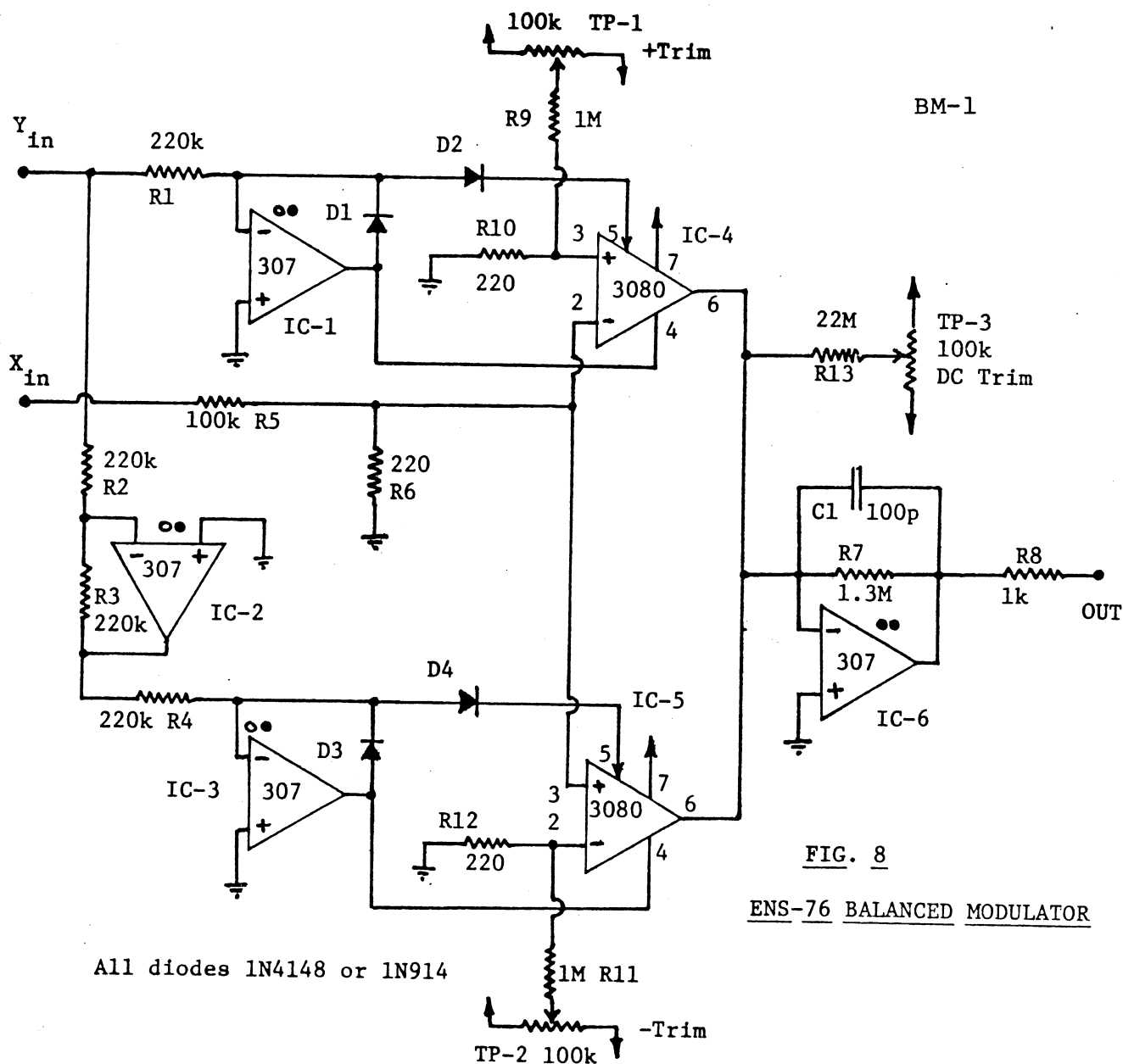
| <u>Preferred<br/>Circuit<br/>Option</u> | <u>Original<br/>Source</u> | <u>Original<br/>Designer</u> | <u>Estimated<br/>Parts<br/>Cost</u> | <u>Notes</u>  |
|-----------------------------------------|----------------------------|------------------------------|-------------------------------------|---------------|
| BM-1                                    | EN#63                      | Hutchins                     | \$5                                 | ENS-76 Series |
| BM-2                                    | MEH 5f (3)                 | *Hutchins                    | \$7                                 | Uses 595      |
| BM-3                                    | EN#134                     | Hutchins                     | \$14                                | Uses AD533    |
| BM-4                                    | EN#113                     | Fritz                        | \$5                                 | Uses LM13600  |

\*straightforward design from application notes

## ENS-76 BALANCED MODULATOR - OPTION 1

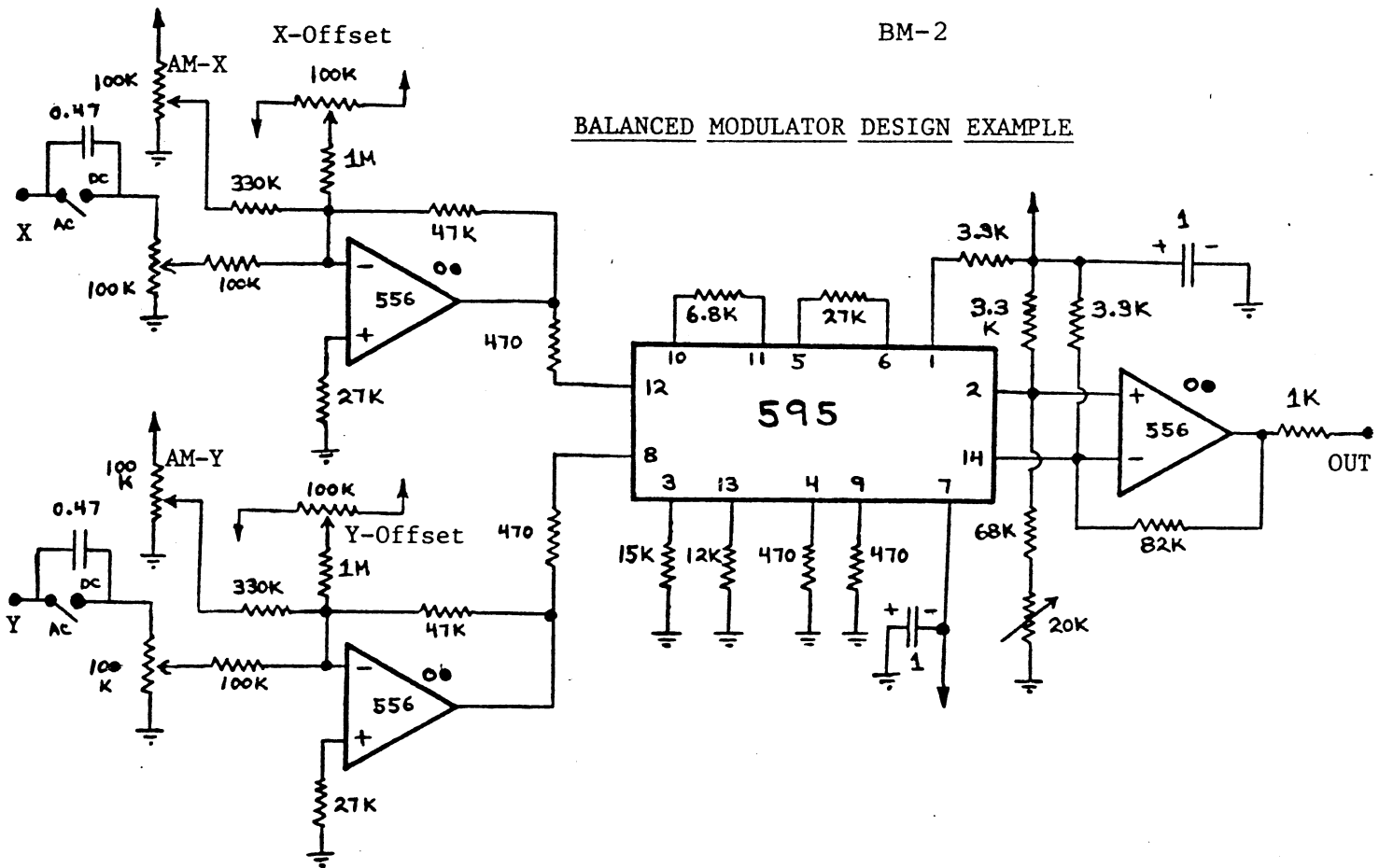
All of our balanced modulators ("Ring" modulators) to date have been based on the type 595 4-quadrant multiplier IC's. These have been satisfactory, although not outstanding (due to insufficient carrier rejection). Also, these have been a little difficult to balance. The new circuit given here may have some performance advantages (the one we built has better carrier rejection than the older 595 design). However, we have another important reason for changing the design - the 595 is getting hard to find. A year or so ago these were available from a number of sources for a price in the range of \$2 to \$4. Currently, we don't know where to get these except from a regular distributor. This means that the builder has to pay a price in a range where he has to seriously consider one of the newer and better 4QM IC's available. The use of one of these newer 4QM's as the heart of a balanced modulator design is an attractive design option. Here however, we are interested in a design that will cost only a few dollars for the IC's. The circuit is shown in Fig. 8.

The present design is basically an extension of the VCA design described above. We have simply connected two 2QM's in parallel. One of the 2QM's is formed by IC-1 and IC-4 while the other is formed from IC-3 and IC-5. The outputs of the CA3080's are currents so we can sum them by just shorting them together. This summed current is applied to I/V converter IC-6. IC-2 is a simple inverter for the Y input (unipolar) of the lower 2QM. Since Y is unipolar (must be positive in this case), only one of the 2QM's is on at any one time. If the Y input is positive, IC-4 supplies current while IC-5 supplies none. The



## DESIGN EXAMPLE

The specific features of this design example have been discussed above. The balancing technique is to first return the AM controls to zero. Next, insert a  $\pm 5$  volt 1000 Hz signal (or so) into the X input (AC coupling with the gain all the way up). Adjust the Y-Offset for minimum signal at the output. Repeat for the Y input. Repeat both steps until no improvement is noted. Balance the DC with the 20k pot, adjusting the 68k resistor up or down if necessary.



The above design example shows full features on both inputs. In some cases, it may be necessary to simplify and just give full features on one input, and a simple input on the other.

use of the (-) input of IC-4 and the inversion due to IC-6 result in a polarity of the output of IC-6 which is the same as the polarity of the X input. If on the other hand, Y is negative, then IC-4 is cut off while IC-5 is supplying current. Since the X input is applied to the (+) input of IC-5, and IC-6 is inverting, the output of IC-6 is inverted with respect to the polarity of the X input. This is how a four-quadrant multiplier is achieved.

From this point on, the rest of the circuit should be self-explanatory if the earlier VCA circuit was understood. A few final points on the design should be made. We used 220k resistors on the Y input so that the actual input impedance would be 100k or greater. This means that R7 had to be approximately twice the corresponding value in the VCA case. (We actually have a 1M and a 270k in series for R7). For best results R1, R2, R3, and R4 should be matched to 1%. The capacitor C1 is needed to remove a glitch that occurs at the zero crossings of the  $Y_{in}$  voltage. If it is desired that the circuit operate above 10 kHz, better performance will be obtained by using type 556 op amps (MC1456 or equivalent) instead of type 307's. Type 556 op-amps can also be used in the VCA circuit. It is (as always) good practice to add power supply bypass capacitors of about 0.02 mfd at various points in the circuit. In particular, these are useful on the supply lines to IC-1 and IC-3.

#### BALANCING THE BALANCED MODULATOR

Balancing the 4QM in Fig. 8 is not difficult. A number of techniques can be used. We suggest DC balancing using a digital voltmeter or a sensitive DC scope. Note that a very accurate DC zero can be obtained by using the scope on a sensitive range and using ground as the zero reference. First you set the level as near zero as possible. Then you short the output point to ground and see if the scope trace deflects. When you can't see it move as the ground is made, you have an accurate zero. Balance as follows:

First with no inputs connected, adjust TP-3 so that the output voltage is zero. Then apply +5 volts to the Y input. If you don't have a +5 source handy, use a +15 source and a 220k series resistor. This puts about +5 volts at the input. This +5 volt signal turns IC-4 on full. Now adjust TP-1 for a zero at the output of IC-6. Next apply -5 volts to the Y input (or -15 through 220k as before). This turns IC-5 on full. Now adjust TP-2 for a zero at the output of IC-6. Recheck the whole procedure.

Now as a check for balance and proper operation, connect +5 to both the X and the Y inputs. Note that if you use the +15 source and the 220k series resistor, you should use two resistors, one for each input. Better still, use a 100k resistor to +15 with the two inputs shorted. The output voltage should be roughly +5 (in the range +4 to +6 say). Now connect -5 volts to both inputs (or -15 through series resistors as with the +15 source). The output should be the same as with +5 on both inputs. Finally, try the combinations of +5 on one input and -5 on the other. This should give the same magnitude as before (the roughly 5 volt level), but the output will be negative. All these values should be within about 10% of each other in magnitude. Average them. If the average is not 5, then you can adjust R7 accordingly. Do not be concerned if the agreement is not better than about 10% unless you are testing with precision resistors and precision voltage sources. It is the DC balance that was done initially that is important, not the exact agreement of the voltage products. Once all this checks out, and you adjust R7 so that you have a 5 volt output for 5 volt inputs, the device should be suitable for most musical applications.

\* \* \* \* \*

-by Bernie Hutchins

A balanced modulator is basically a four-quadrant multiplier. Thus the design of a balanced or "ring" modulator is a matter of selecting a suitable four-quadrant multiplier chip and setting up the necessary input/output and trim circuitry. There are a good number of suitable multiplier IC's, including the 595 series which have been around for more than 10 years. We have also seen designs based on various OTA devices.

In actually applying a designed unit, we find that for musical purposes, a couple of secondary problems perhaps are more important than the initial quality of the multiplier IC used. These problems arise mainly as a result of the use of the balanced modulator as a modular unit. First, consider that we may have a balanced modulator trimmed perfectly, and then put it in our system. We plug in a couple of signals, and there is a carrier feedthrough heard. The reason is simple - the signals have some DC offset of their own, and this of course upsets the balance. This suggests AC coupling, but we find this a problem as well, as we may want to multiply an AC signal by an envelope which will be blocked by AC coupling. Thus the only practical solution would seem to be to provide switchable AC or DC coupling, separate for each channel of the input.

The second problem relating to modularity is that a module may be built, tested, and balanced on one power supply system and then installed elsewhere, requiring a rebalance. Related problems are found where the supply the modulator is working off is required to supply varying loads, as when a new module is added to the system. This may shift a supply voltage a small amount, but enough to throw off the trimming section. The solution to this problem is to supply some sort of local regulation to the module itself.

The figure on the next page shows a circuit using the AD533 multiplier chip (Analog Devices, Norwood, MA), which implements switched AC/DC coupling and local on-board regulation. The same principles could be used with other multiplier chips, but we have become rather fond of the AD533 since it is so easy to use and trims so well.

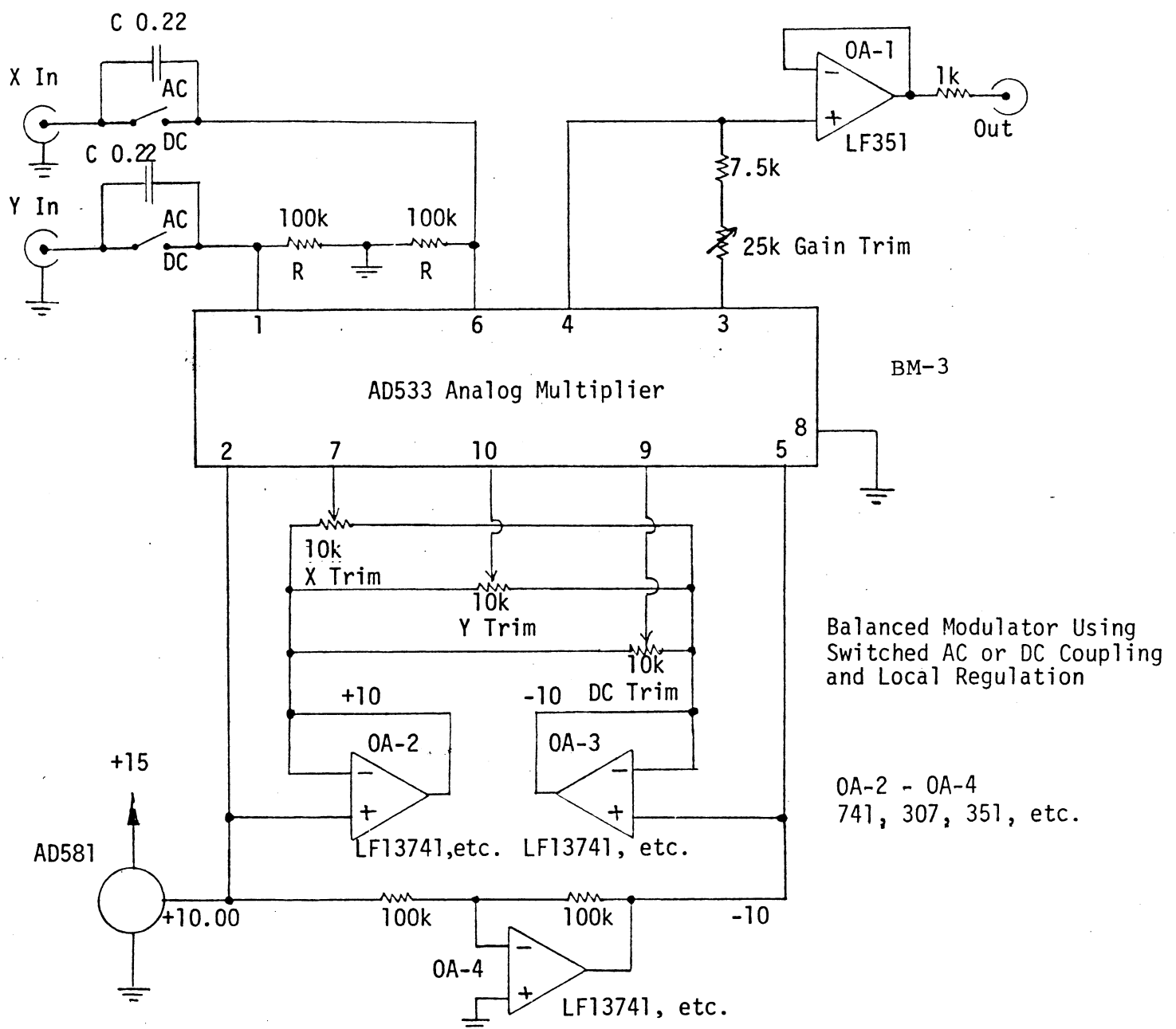
AC coupling is achieved using the capacitors on the X and Y input lines as shown. With the values shown,  $R = 100k$  and  $C = 0.22$  mfd, we calculate a cutoff frequency for AC coupling of:

$$f_c = 1/2\pi RC = 7 \text{ Hz}$$

so we can be pretty confident of DC rejection while at the same time, audio frequencies above 15 Hz will be passed through. To defeat the AC coupling, it is only necessary to short the capacitors with the switches shown. For normal "ring" modulator effects, AC coupling would be used on both inputs, as both signals would be audio. To use the module as a VCA, one of the inputs would be AC coupled and attached to the signal to be controlled. The other would be DC coupled and attached to the controlling envelope.

The output section is given its own op-amp buffer OA-1 even though pin 4 of the 533 is the output of an op-amp internal to the 533. We do this so that power requirements to the 533 will be fairly constant, and not depend on the external load. Note that all op-amps shown here run on the usual +15 and -15 supplies.

The portion of the circuit below the AD533 is concerned with on-board regulation. The heart is the AD581 precision +10.00 volt reference. Here it is the stability and not the precision that we need, but the AD581 is ideal in this area too. We will be running the AD533 from a  $\pm 10$  volt supply, the AD581 supplying +10 and OA-4, the op-amp inverter, supplying the -10. Since the 533 requires about 6 ma current, and the 581 can provide 10 ma, this should be a safe system. However, so as not to tax the current capability of the 581 or of OA-4, additional buffers OA-2 and OA-3 are used to supply current for the trimmers which require 2 ma each. Note that the system here is intended for signal levels of  $\pm 5$  volts.



Trimming the circuit is very easy. First switch to AC coupling. With no inputs attached, measure the output DC voltage, and adjust DC-Trim until this voltage goes to zero. Next, attach an AC signal (say 100 Hz to 1000 Hz) to the X input, and adjust the Y-Trim for minimum signal through to the output. Then attach the AC signal to only the Y input, and adjust Y-Trim for minimum signal at the output. Then set the X input to +5 volts DC, and switch it to DC coupling, leaving the AC signal connected to the Y input. Adjust the "Gain Trim" so that you have the same level signal at the Y input and at the output. Recheck all these steps, and make finer adjustments if necessary. The board is now ready to use.

In all our work with the AD533 multiplier we have found it very easy to trim, and most of the trimmers are set almost at their centers. This suggests a simple modification where resistor dividers are used at the inputs of OA-2 and OA-3 to lower these voltages to say +2 and -2 volts. This would make the trimmers even easier to use here.

This example has given the general principles of using an analog multiplier as a balanced modulator module, and has given a circuit example which is now to become our preferred circuit for this function.

LM/XR 13600 TRANSCONDUCTANCE AMPLIFIER:

-by Ian Fritz

I. INTRODUCTION

The module described in this article serves as both a VCA (voltage controlled amplifier, or two quadrant multiplier, or 2QM) and as a balanced modulator (four quadrant multiplier or 4QM), with the mode of operation selected by a switch. The design has evolved from previous designs using transconductance amplifiers, so much of the circuitry will be familiar to Electronotes readers. New features of the design include the following: 1) The new 13600 dual transconductance amplifier [1] available from National Semiconductor and Exar is used for improved signal-to-noise performance ( ~10 db improvement) 2) A modification of the simple 4QM design with a single transconductance amplifier - which is given in the manufacturers' application notes [2] and which was discussed by Hutchins [3] - is employed. This modified 4QM design has better accuracy than the original circuit. 3) The design is efficient in that it only used 1-1/2 chips: a quad op-amp plus half of a 13600 4) Finally, the voltage-to-current converter for the control input is somewhat different from previous designs. A "bias" control that selects a weighted average of the Y input and a fixed bias voltage replaces the usual "initial gain" control of previous VCA designs, and also serves to unbalance the input in the 4QM mode to mix in the carrier signal.

II. DESIGN OF THE TRANSCONDUCTANCE SECTION (X INPUT)

To begin with, it will be assumed that the amplifier bias current  $I_b$  driving the 13600 is given by the following expressions, where  $V_y$  is the control voltage, or the Y input voltage:

$$I_b = \frac{V_y + 5}{20} \quad (4QM) \quad (1a)$$

$$I_b = \frac{V_y}{10} \quad (2QM) \quad (1b)$$

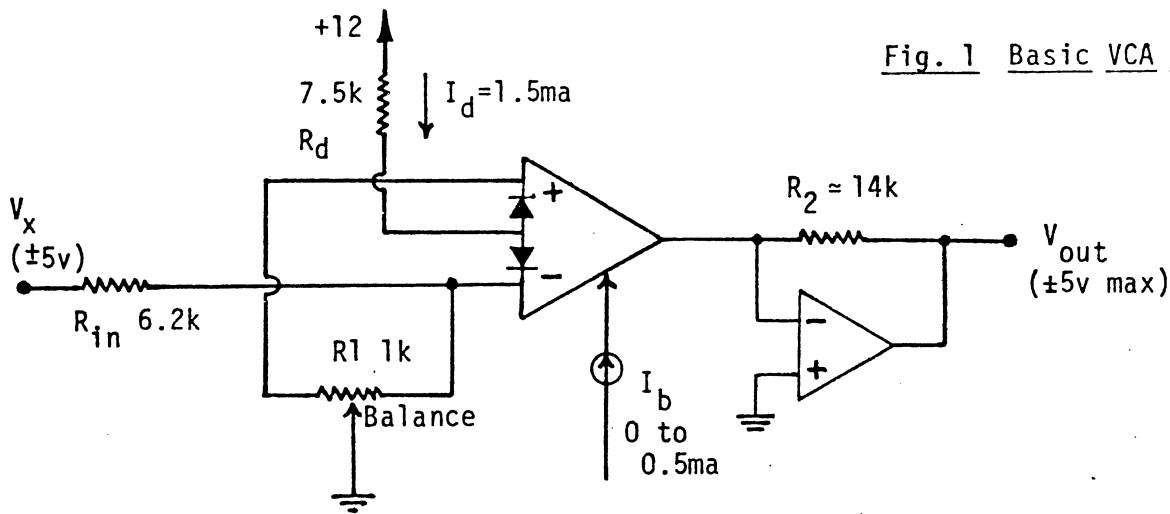
For the 4QM case  $V_y$  is between -5 and +5 volts, and in the 2QM case  $V_y$  is between 0 and +5 volts. Thus  $I_b$  covers the range 0 to 0.5 ma in both cases. Design of the circuitry to produce  $I_b$  will be discussed in Section III.

A fairly extensive study, both theoretical and experimental, of the operation of the 13600 has been undertaken. The details of this investigation will not be given here, but the major results will be briefly summarized. First it must be pointed out that the analysis used by Hutchins [1] is not quite correct. This is because the current through the input resistor of Figure 6b of Reference 1 is (approximately) equal to twice the current  $I_s$  of Figure 3 (same Reference). The assumption that these currents are equal is what leads to the measured output voltage appearing to be off by a factor of two.

It must be emphasized that the diode predistortion available with the 13600 does not provide exact compensation of the input nonlinearities, as does the input structure of the SSM chips [4]. The manufacturers only claim about a 10 db improvement in signal-to-noise ratio (referred to 0.5% harmonic distortion), and it does not appear possible to do much better. This means that the inputs can be driven to about  $\pm 30$  mv as opposed to the  $\pm 10$  mv used without predistortion. A design utilizing  $\pm 50$ mv of drive was developed, but it required an active current source for the diode bias, and had only about 10% headroom before the onset of hard limiting.

It is also found that as large a diode bias  $I_d$  as possible should always be used. The suggestion [1] of increasing the input drive by decreasing  $I_d$  is not recommended, as this increases the nonlinearities of the input structure. The present design uses  $I_d = 1.5$  ma. The absolute maximum allowed value is 2 ma.

Fig. 1 Basic VCA Structure

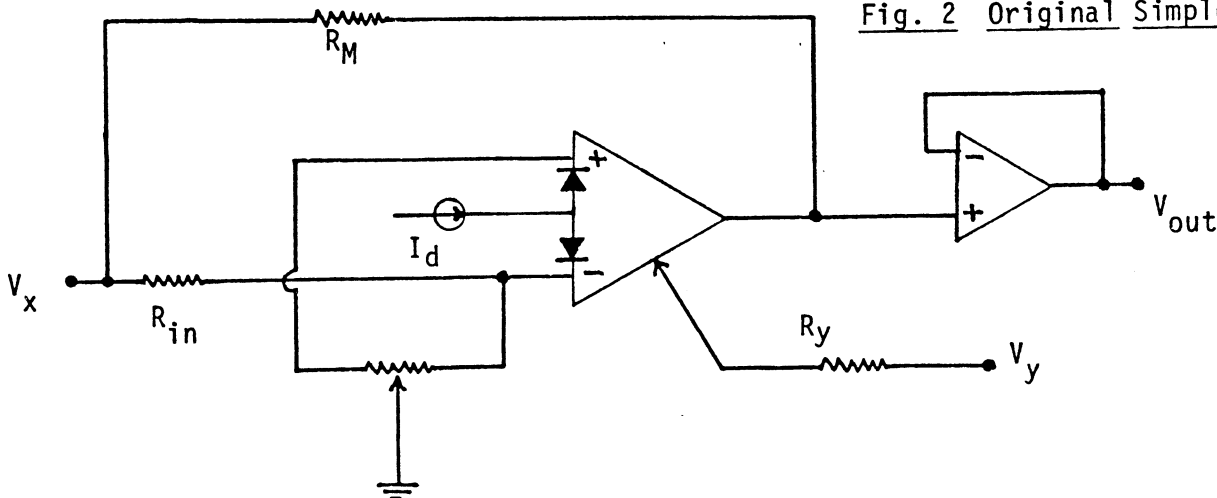


The basic circuit for the 2QM is shown in Figure 1. The value of  $R_{in}$  was determined by solving the equations for the input network, and  $R_4$  was determined empirically to give unity gain. It was found that  $R_4$  was smaller than predicted, possibly due to non-ideal transconductance of the chip (a range of about  $\pm 30\%$  is specified by the manufacturers).

As mentioned, the 4QM is a variation of the circuit described previously by Hutchins [4]. That circuit is shown in Figure 2. Analysis of Figure 2 shows that the overall gain is determined by the ratio of  $R_M$  to  $R_Y$ , which in turn is determined by the balance condition for the multiplier. The situation is different if the setup of Figure 3a is used. Here the Y voltage input is converted to a current by active circuitry. The VCA design corresponding to Figure 3a is given in Figure 3b. A fairly simple analysis yields the following neat result: For the 2QM, assume that  $R_{out}$  is chosen for unity gain at maximum bias current, so that  $V_{out} = V_x V_y / 5$ . Then the 4QM will be properly balanced and have  $V_{out} = V_x V_y / 5$  when  $R_M = 2 R_{out}$ . Now it is easy to see that a suitable 2QM/4QM could be made based on Figure 3: the bottom end of  $R_{out}$  could be switched either to ground (2QM) or to a resistor whose other end is connected to  $V_x$  (4QM). Readers who have been following this rather sketchy argument carefully will have noticed that the two V-I converters in Figure 3 are different, so some switching is required there also.

A problem, whose origin is not understood, arises with the circuits of Figure 3. Both circuits are quite nonlinear as a function of  $I_b$ . This nonlinearity ( $\sim 20\%$ ) does not appear to depend on the value of  $V_x$  ( $V_x$  is fixed and  $I_b$  is varied), nor on the value of  $I_d$ . The nonlinearity does not occur, however for the VCA of Figure 1, and therefore appears to be associated with the output structure. Fortunately, there is a 4QM structure analogous to Figure 1, and this is shown in Figure 4. The second op-amp in this circuit is interesting in that it simultaneously works as a current to voltage converter for the output of the 13600

Fig. 2 Original Simple 4QM





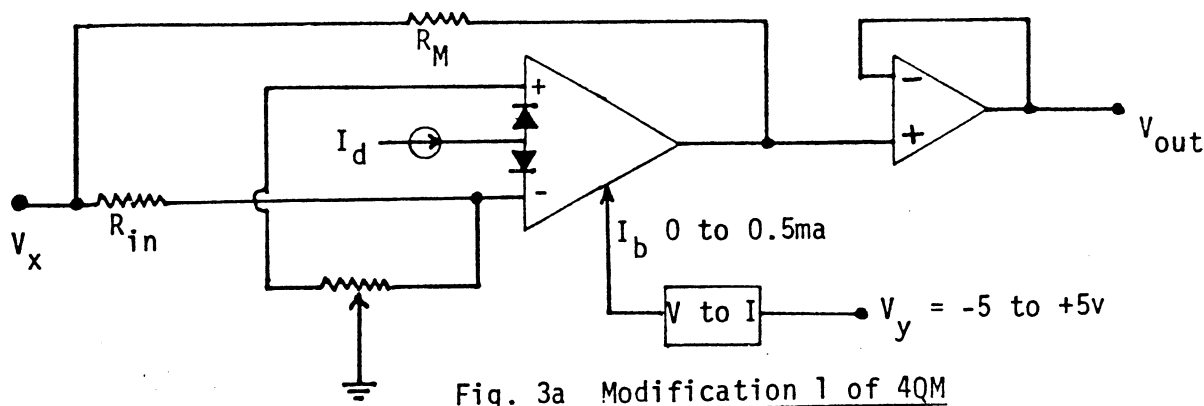


Fig. 3a Modification 1 of 4QM

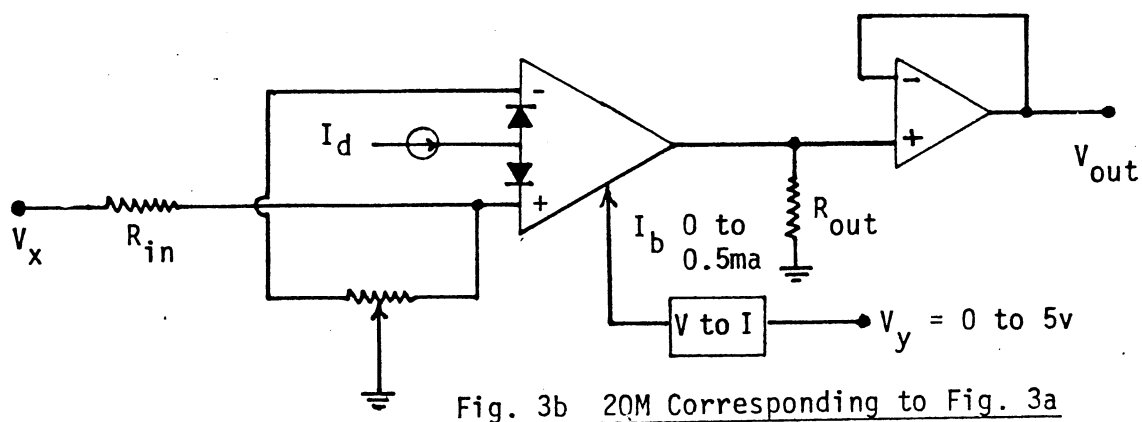


Fig. 3b 2QM Corresponding to Fig. 3a

and as an inverting amplifier for the input voltage. Analysis indicates that the 4QM is balanced and has  $V_{out} = V_x V_y / 5$  when  $R_a = R_b = 2R_2$ , where  $R_2$  is shown in Figure 1. Thus we can change the 4QM of Figure 4 to the 2QM of Figure 1 simply by moving the left-hand end of  $R_a$  from the input of the circuit to the output! (Again, of course, we have to change the V-I converter at the same time).

### III. CONTROL SECTION (Y INPUT)

For the Y input circuitry it is necessary both to implement Eqns. 1 and to provide some sort of manual gain control. A means to unbalance the 4QM is also a useful feature. The design adopted here is shown in Figure 5. The two resistors  $R_a$  and  $R_b$  are equal. It is convenient to analyze the structure from the end backwards. The V-I converter produces an output current  $I_b$  ranging from 0 to 0.5 ma in response to an input  $V_a$  ranging from 0 to -5 volts. If the control  $R_b$  is in the CW position, then  $V_a$  is -5 volts, independent of  $V_y$ .

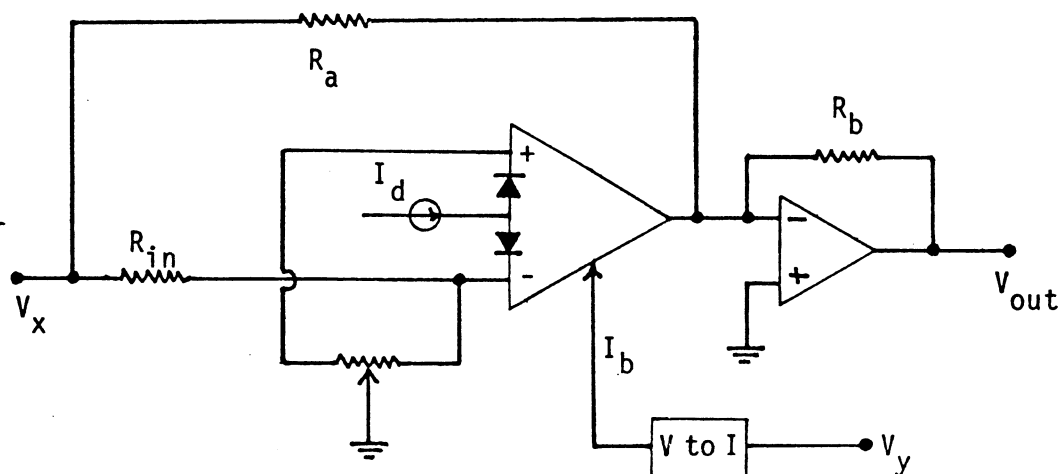


Fig. 4 Modification 2 of 4QM for Improved Accuracy

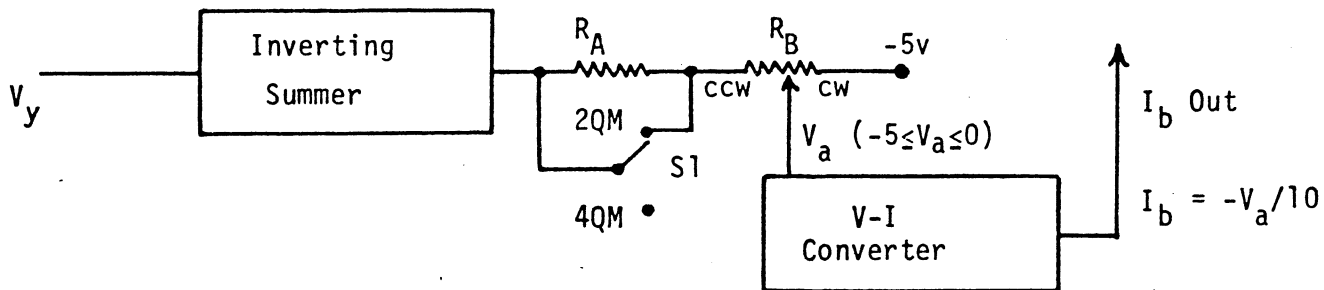


Fig. 5 Schematic of Control (Y) Section

Thus the transconductor is turned on to full gain and the output of the module is  $V_x$ . (This statement is true independent of whether the X section is in the 2QM or 4QM mode). Now suppose  $R_B$  is in the CCW position. There are two cases to examine. First suppose that the switch S1 is closed (2QM position) and the transconductor is in the 2QM mode (Figure 1), so that the unit is being used as a VCA. Values of  $V_y$  from 0 to 5 volts are of interest, so  $V_a$  clearly ranges from 0 to -5 volts. Thus the transfer function of Figure 5 is

$$I_b = \frac{V_y}{10} \quad (2)$$

as required by Equation 1a. The second case is for 4QM operation with S1 open. Here  $V_y$  ranging from -5 to +5 is of interest. Since  $R_A$  and  $R_B$  are equal,  $V_a$  is the average of  $-V_y$  and -5 v. Thus  $V_a = (1/2)(-V_y - 5)$ , and

$$I_b = -\frac{V_a}{10} = \frac{V_y + 5}{20}, \quad (3)$$

as required by Equation 1b.

In general  $R_B$  provides a weighted average of  $V_y$  and -5 volts that allows the modulation or control input voltage to be "traded off" against a fixed bias. The advantage of this kind of control is that it is normalized in the sense that  $I_b$  is always restricted to its proper operating range.

#### IV. FINAL CIRCUIT

The final circuit is given in Figure 6. A couple of points need to be made here. A non-inverting buffer is used on the X input (OA1a) to preserve signal polarity through the unit. An inverting summer could be used if the overall inversion does not matter. The purpose of R9 is to offset the Y balance slightly when PC3 is in the full CCW position. This allows the control envelope to be burried slightly in the 2QM mode, and gives some range for the Y balance point in the 4QM mode. The correct position for PC3 can generally be set by ear. The circuit as drawn is set up for  $\pm 12$  v supplies. For  $\pm 15$  v operation R4 should be changed to 10 k $\Omega$  and R13 to 2 k $\Omega$ .

The circuit as indicted has a bit of distortion which can be seen on a scope, but which doesn't seem to have any significant audible effect. The circuit has smooth limiting and saturates at about  $\pm 8$  volts. Predicted values for  $R_a$  and  $R_b$  (Figure 4) are 38 k $\Omega$ , whereas the measured values are about 33 k $\Omega$  and 26 k $\Omega$ , respectively. The gains of the 2QM and the 4QM are slightly different (within about 15%), and TP2 may be set to give exact unity gain for one or the other, but not both. The setup procedure for the unit is as follows:

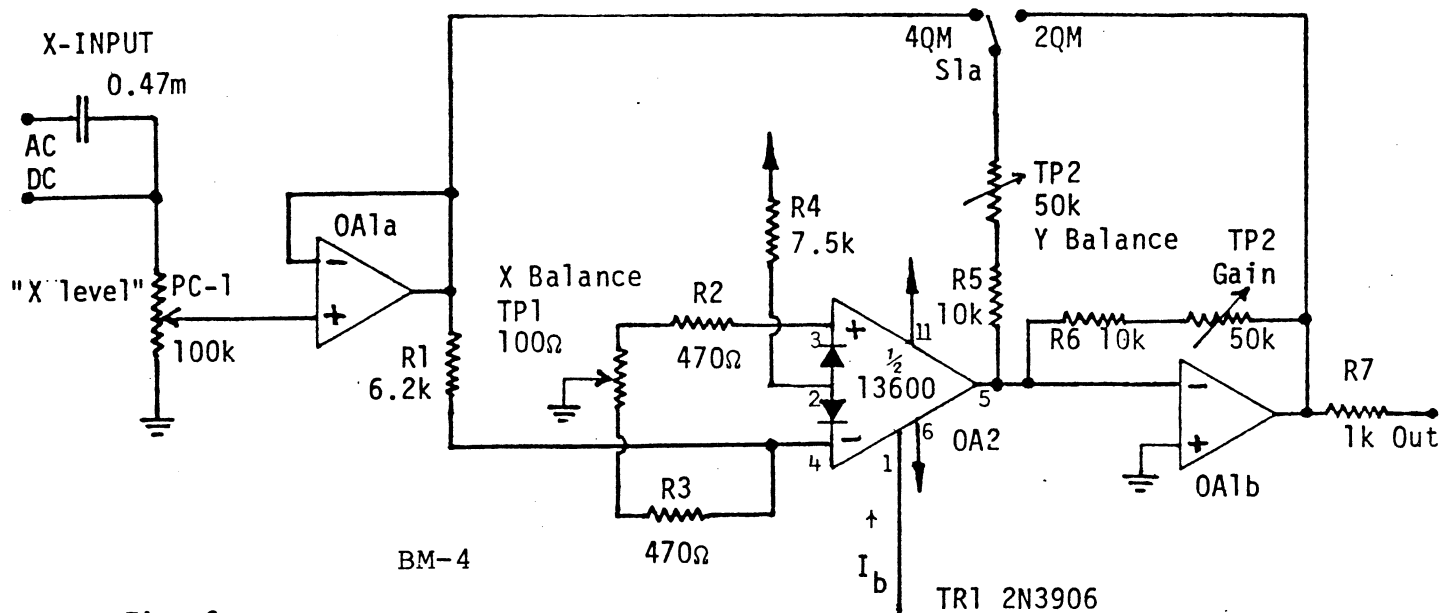
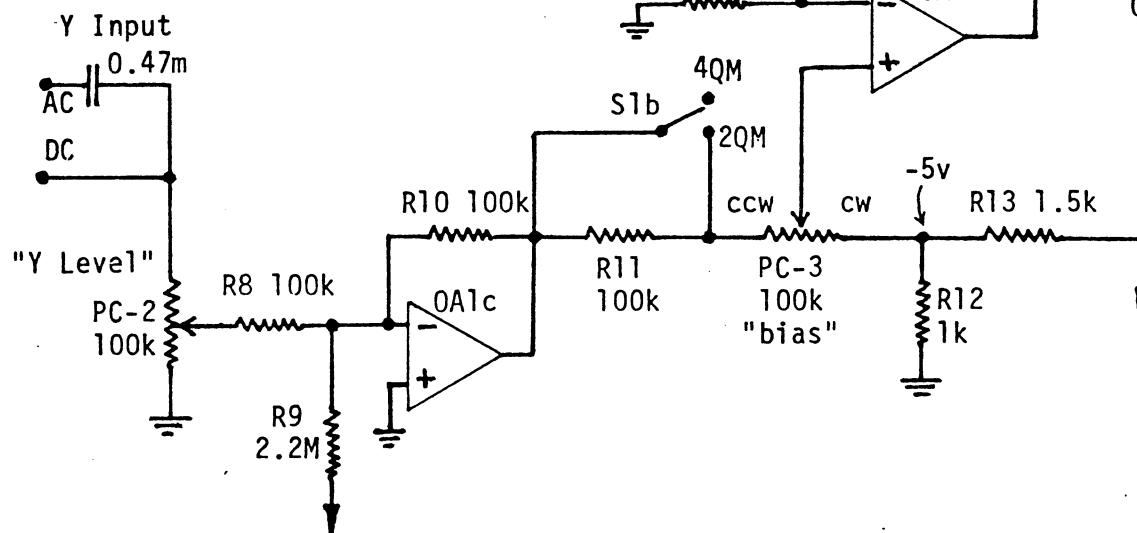


Fig. 6

Switched 2QM/4QM Module



OA1 = Quad BiFet  
e.g. TL074  
or TL084

OA2 = 1/2  
LM/XR13600

- Step 1. S1 position ("mode"): 2QM  
PC3 ("bias" control): full CCW  
X input: 0 v  
Y input:  $\pm 5$  v audio, or + 5 v pulse  
Adjust TP1 (X balance) for minimum feedthrough
- Step 2. S1: 2QM  
X input:  $\pm 5$  v audio  
Y input: 0 v  
Adjust "bias" control (PC3) to just below point where VCA starts to turn on. This is the "threshold" position and should be marked on the panel. This position represents the nominally correct Y balance position. Leave PC3 in this position for the remaining adjustments.
- Step 3. S1: 4QM  
PC3: "threshold"  
X input:  $\pm 5$  v audio  
Y input: 0  
Adjust TP2 "Y balance" for minimum feedthrough

- Step 4. S1: 4QM or 2QM  
PC3: "threshold"  
X input:  $\pm 5$  v audio  
Y input: + 5 v dc  
Adjust TP2 "gain" for unity gain
- Step 5. S1: opposite from Step 4  
X input:  $\pm 5$  v audio  
Y input: + 5 v dc  
Check for approximately unity gain.

#### REFERENCES

1. B.A. Hutchins, Electronotes 107, 3 (1979).
2. Exar Integrated Systems, 750 Palomar Ave., P.O. Box 62229, Sunnyvale, CA 94088, Application notes for the XR-13600.
3. B.A. Hutchins, Electronotes 107, 13 (1979).
4. D. Rossum, Electronotes 67, 3 (1976).

\* \* \* \* \*

## FREQUENCY SHIFTER

Frequency Shifter: This collection contains only one frequency shifter, and is presented in the form of a reprint of the original design in EN#83



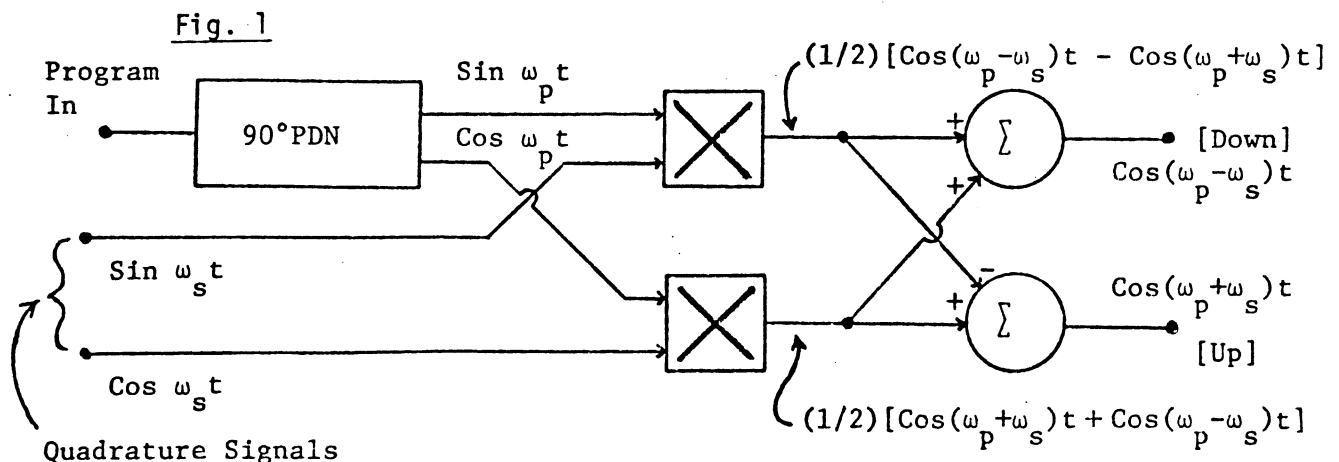
# THE ENS-76 HOME-BUILT SYNTHESIZER SYSTEM - PART 9, FREQUENCY SHIFTER:

-by Bernie Hutchins, ELECTRONOTES

**INTRODUCTION:** Basically, there is no new technology to be presented in this part of the ENS-76 series. The design ideas have been around for some time [1-3] and we have presented a full theory of frequency shifter design [4]. The purpose here is to clean up the design a little, using some new suggestions [5-7]. The frequency shifter is a module that is available from only a couple of manufacturers [8], so we feel it is important that our home-built version be as up-to-date as possible. We shall attempt here to present only a very brief discussion of the operating theory, and to confine our presentation to construction details. A full discussion of the theory along with the original design example can be found in reference [4].

## BASIC THEORY:

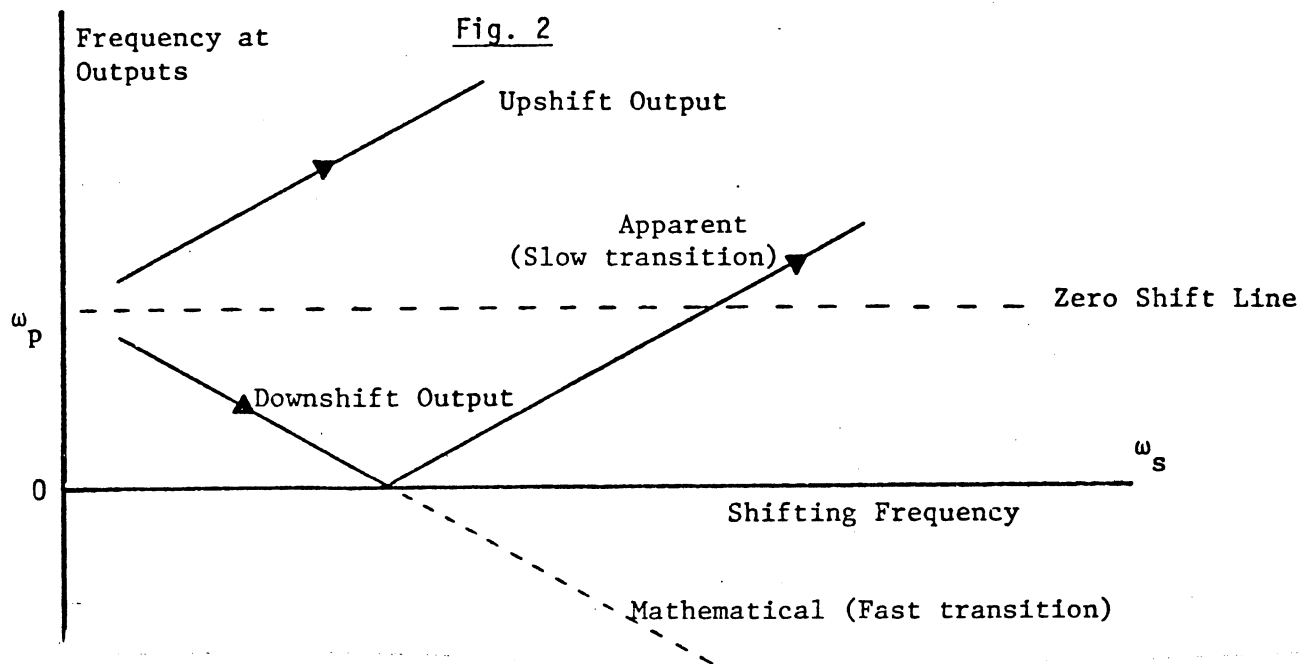
Although there are several means of achieving a frequency shift, we will be using the scheme shown in Fig. 1. Basically the idea is to form quadrature (Sine and Cosine) components of the signal to be shifted and of the signal that does the shifting. In the event that these signals are complex, quadrature pairs are formed component by component. In Fig. 1, it is assumed that only the program is complex. The quadrature splitting of the program material is accomplished by means of a wide-band 90° Phase Difference Network (90°PDN). The 90°PDN splits a sinusoidal component within its band of operation into signals having phases  $\phi$  and  $\phi+90^\circ$  where  $\phi$  is some angle that depends on the frequency. In practice, the splitting is not exactly 90° but will vary about 90° within some limit that is typically on the order of 1°. Note that the waveshape of signals passing through the 90°PDN is not preserved. However the phase-shift circuits in the 90°PDN are all-pass, so the spectrum is preserved. It is this spectrum that we are interested in shifting. The expressions for the various signals in the diagram show how a typical sinusoidal component of frequency  $\omega_p$  is processed. Note that both upshifted and downshifted versions are available by appropriately multiplying and summing the quadrature signals. It is important to realize that all component frequencies in the program signal are shifted by the same number of Hertz, not by the same musical interval. Thus some dramatic changes of tone color can result.



The only parts of the design that are not reasonably straightforward are the 90°PDN and the quadrature oscillator that provides the shifting signal  $\omega_s$ . Of course, it is possible to use an ordinary sine wave oscillator and a second 90°PDN to provide the shifting signal. This allows us to input complex shifting signals as well. However, at times we want very small shifting frequencies, and it is easier to construct a wide range quadrature oscillator than it is to extend the range of the 90°PDN down to very low frequencies. Note however that the scheme of Fig. 1 is independent of

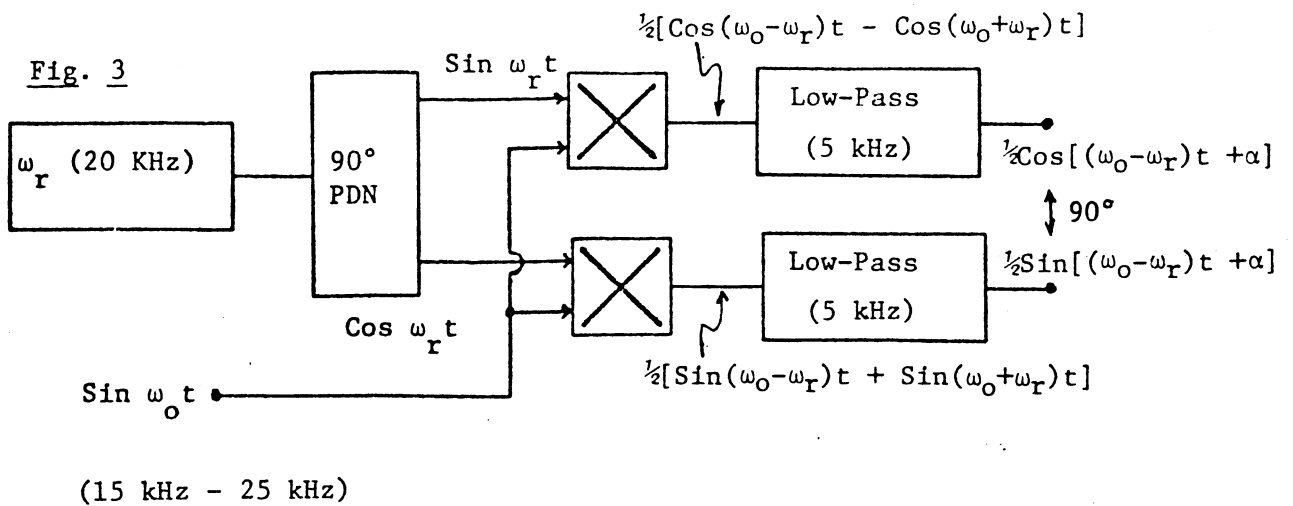
the means by which the quadrature signals are produced, and it is useful to employ other quadrature devices such as beat-frequency quadrature oscillators as these can be used to provide shifts through zero shift.

Fig. 2 shows how the output frequencies shift as a function of the shifting frequency  $\omega_s$ . Note that in this case neither upshift nor downshift reaches the zero shift line at  $\omega_s = 0$ . This is because we are assuming that our quadrature oscillator won't go all the way to zero. This will be the case with simple quadrature oscillators. Thus, it is not possible to get zero shift as shown. On the other hand, the downshifted signal reaches zero when  $\omega_s = \omega_p$  and then the downshifted signal becomes an upshift. Mathematically the downshifting just keeps right on going through zero to negative frequencies, and is not really an upshift. It may however be heard as an upshift if the transition is slow. This is because in the region between about 15 Hz and -15 Hz the human ear does not hear the signal, and if the transition through this region takes a substantial time (say 200 ms) then the absolute phasing is "forgotten" and a pure upshift is heard. On the other hand, when the transition through the region around zero is quite short, the negative frequency manifests itself as a change of direction of the waveform, and the result is the same as through-zero frequency modulation (see the ENS-76 VCO Option 4 in EN#75 and referenced material). Thus, the frequency shifter even in the simple form of Fig. 1 and Fig. 2 is capable of through zero frequency modulation. The characteristic direction reversal of the waveform during through-zero modulation was observed experimentally with the frequency shifter.

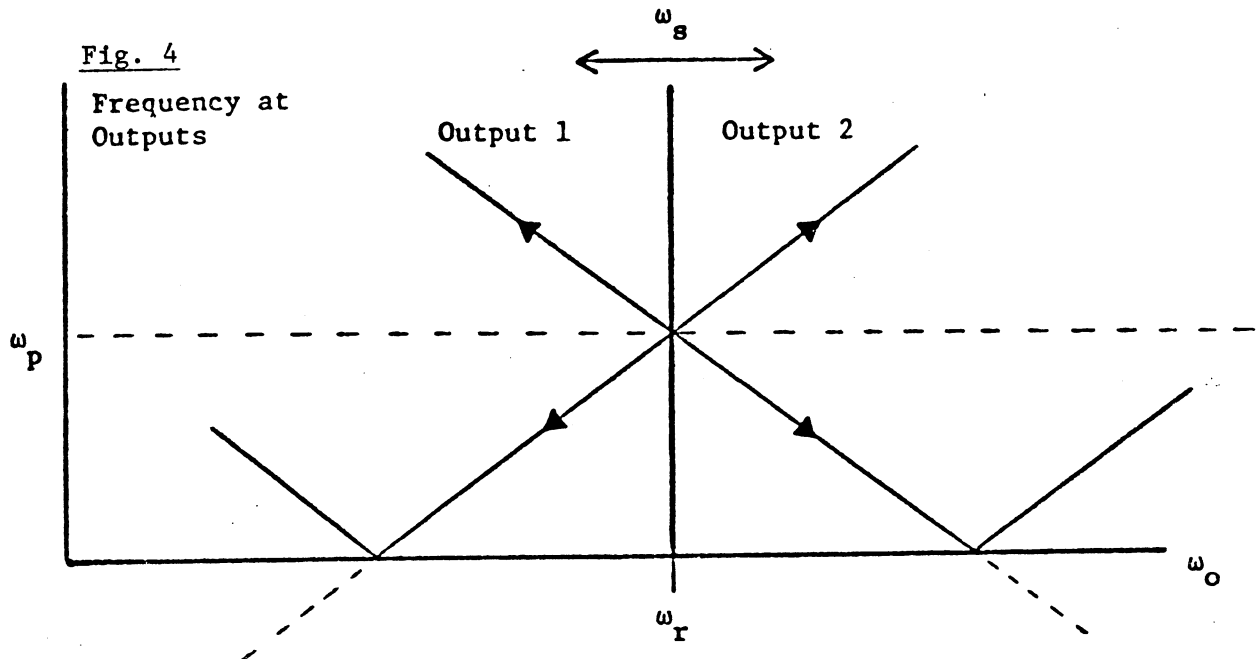


While we could use a simple quadrature oscillator, in this presentation we will be using a beat-frequency method so that the shifter will be capable of shifting through zero shift. The significance of this to frequency modulation and to the processing of live signals will be discussed later. For now, we can just think of the beat-frequency method as another method of obtaining quadrature signals. The beat-frequency method is illustrated in Fig. 3. We start with a sine wave oscillator at about 20 kHz, and use a phase shift circuit adjusted to give a  $90^\circ$  phase shift at 20 kHz. We then beat these two phases against a variable sine wave that varies about 20 kHz. The output of the multipliers contains sum and difference frequencies. The sum frequencies are filtered out by low-pass filtering. Note that the quadrature phase of the two signals is preserved. Bear in mind that the two multipliers shown in Fig. 3 are in addition to the multipliers shown in Fig. 1, so the complete shifter will contain four multipliers. Also, this method also needs a variable oscillator, and a VCO is best. You could plan on using a VCO from your system, or you can build in a VCO as we suggest here.



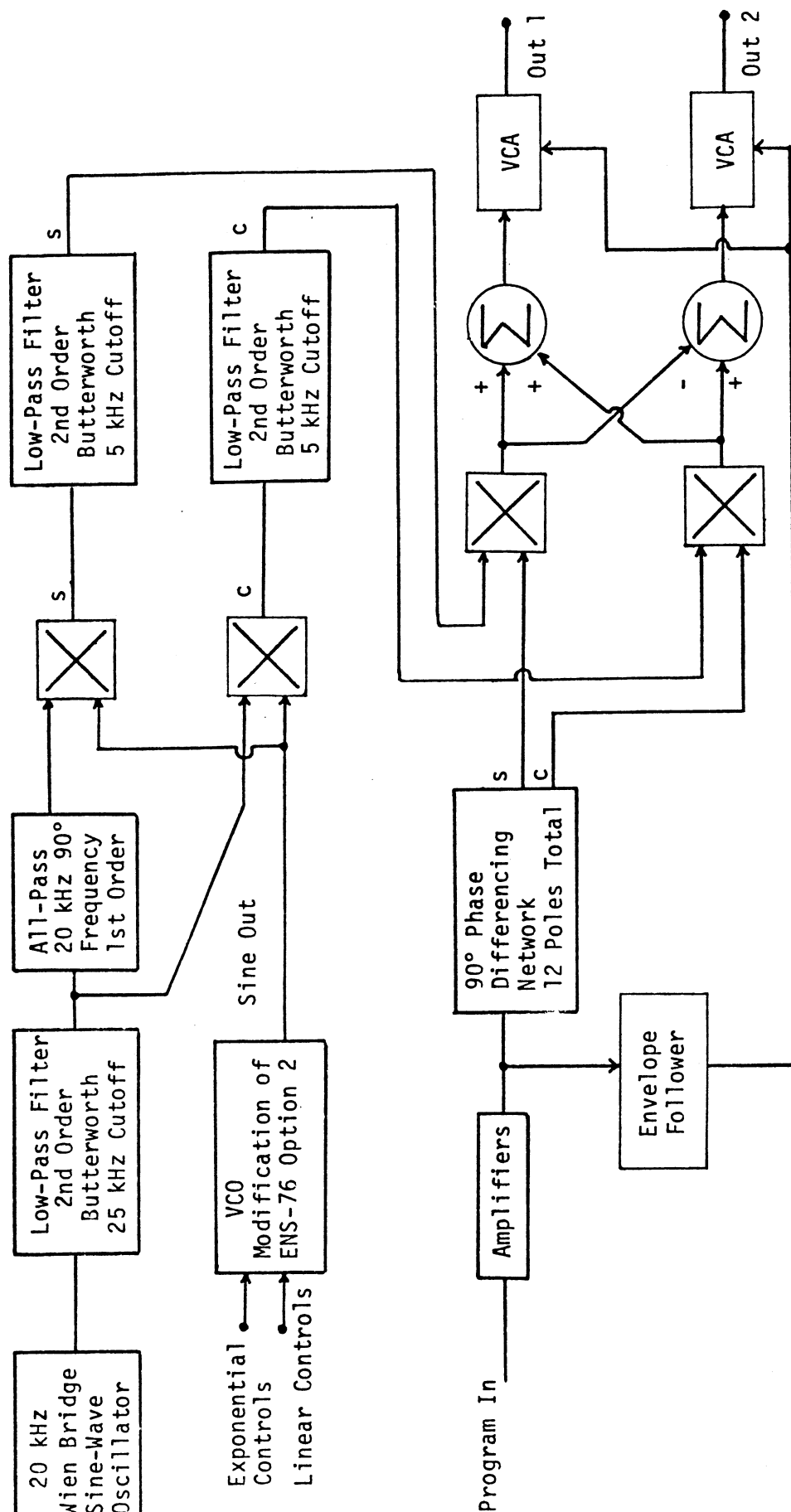


The shifting obtained with the beat-frequency quadrature oscillator is shown in Fig. 4. This can be compared with Fig. 2 which corresponds exactly to the portion of Fig. 4 that is slightly to the right of the  $\omega_r$  line. Note that the outputs cross over so that an upshift becomes a downshift as we pass through zero shift.



#### CONSTRUCTION:

As can be seen from Fig. 1 and Fig. 3, the frequency shifter consists of three major sections. First, there is the quadrature signal source (implied in Fig. 1 and exemplified by Fig. 3). Secondly, there is the 90°PDN. The third major section is the network formed from the multipliers and the two summers in Fig. 1. Note that if we are going to use the beat-frequency method of producing quadrature signals (as we are going to use), the frequency shifter module also needs a VCO to supply a sine wave. We could supply this sine wave from an external VCO, but we will show it as part of the shifter module. Since we will want to process live sounds, it is necessary to have amplifiers to boost signals from microphones and tape recorders to synthesizer levels. Also, keep in mind that for a number of reasons the multipliers in Fig. 1 will not completely block the shifting signal in the absence of a program signal. Thus it is useful to have some additional "squench" circuitry that will block the final output unless some sort of program signal is actually being applied. The squench circuitry consists of a crude envelope follower and a pair of simple VCA's. A full block diagram of the ENS-76 frequency shifter is shown in Fig. 5.



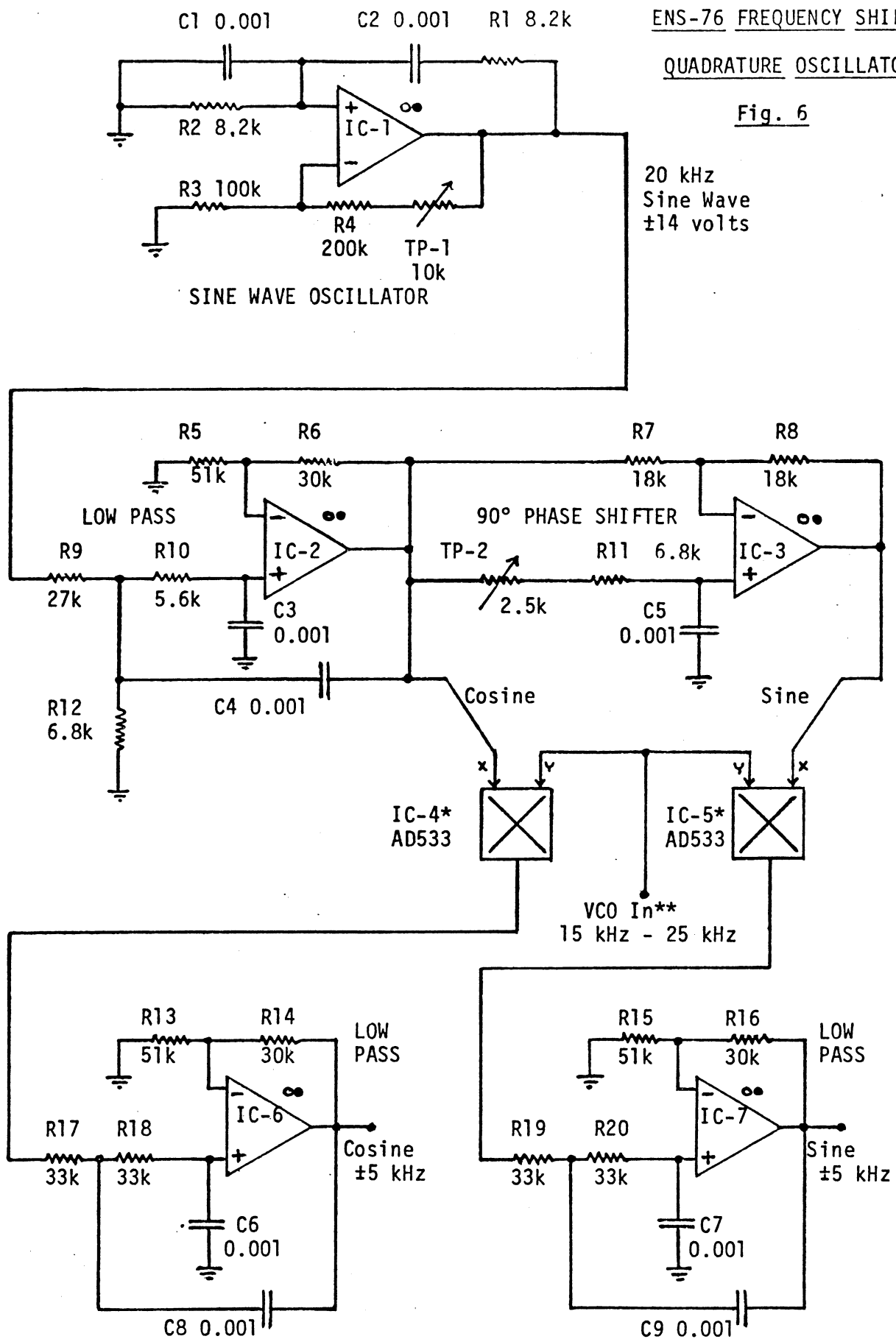
s = "sine phase" relative to corresponding "cosine phase" c

c = "cosine phase" relative to corresponding "sine phase" s

All multipliers formed from type AD533 IC's

## QUADRATURE OSCILLATOR

Fig. 6



\* See Fig. 7

\*\*See Fig. 8

All Op-Amps MC1456, LF351, etc. Do not use 741, 307, etc.

The complete schematic diagram of the frequency shifter is shown in figures 6 through 11. The complete top half of Fig. 5 is realized as in Fig. 6 (with figures 7 and 8 giving additional details on the structures in Fig. 6). Fig. 9 shows the input amplifiers and the envelope extractor. Fig. 10 shows the wide-band 90°PDN that splits the input signal into quadrature components. Fig. 11 shows the two multipliers and summers which form the two outputs, and also the two VCA's which serve to block the output in the absence of a minimum input signal.

The quadrature oscillator shown in Fig. 6 is of the beat-frequency type. IC-1 forms a simple sine wave oscillator of the Wien Bridge type. The nominal frequency of oscillation is 20 kHz, although the frequency is not critical. If you want to trim up the frequency a little, add series resistors of about 100 ohms to both R1 and R2 to decrease the frequency, or shunt resistors of about 100k to R1 and R2 to increase the frequency. Adjust these values as needed. If you are soldering in the components, let them cool at least one minute before taking a reading on frequency. To start up the oscillator, first set TP-1 so that the full 10k is in series with R4. This assures that oscillation will begin. You will note some clipping of the sine wave on the high side. We are using this clipping to control the amplitude, so it is normal. The distortion produced by the clipping will be greatly reduced by the filter formed around IC-2. For final setup, TP-1 is adjusted so that the clipping is reduced. You will note that it is possible to obtain a very nice looking sine wave at the output of IC-1 by lowering the resistance setting of TP-1. However, do not use this setting, as drift may reduce the gain and stop oscillation. It is best that a small amount of clipping be present. The filter formed around IC-2 is simply included to purify the waveform. If you wish, you can consider it to be part of the oscillator. Note that it is essentially a 2nd order Butterworth filter. The attenuator R9:R12 serves to provide about 5.6k impedance (equivalent to R10), and to reduce the amplitude out of IC-1 so that allowing for a gain of 1.6 (R5 and R6) the output of IC-2 is at a ±5 volt level. See application notes AN-7 and AN-37 for more information.

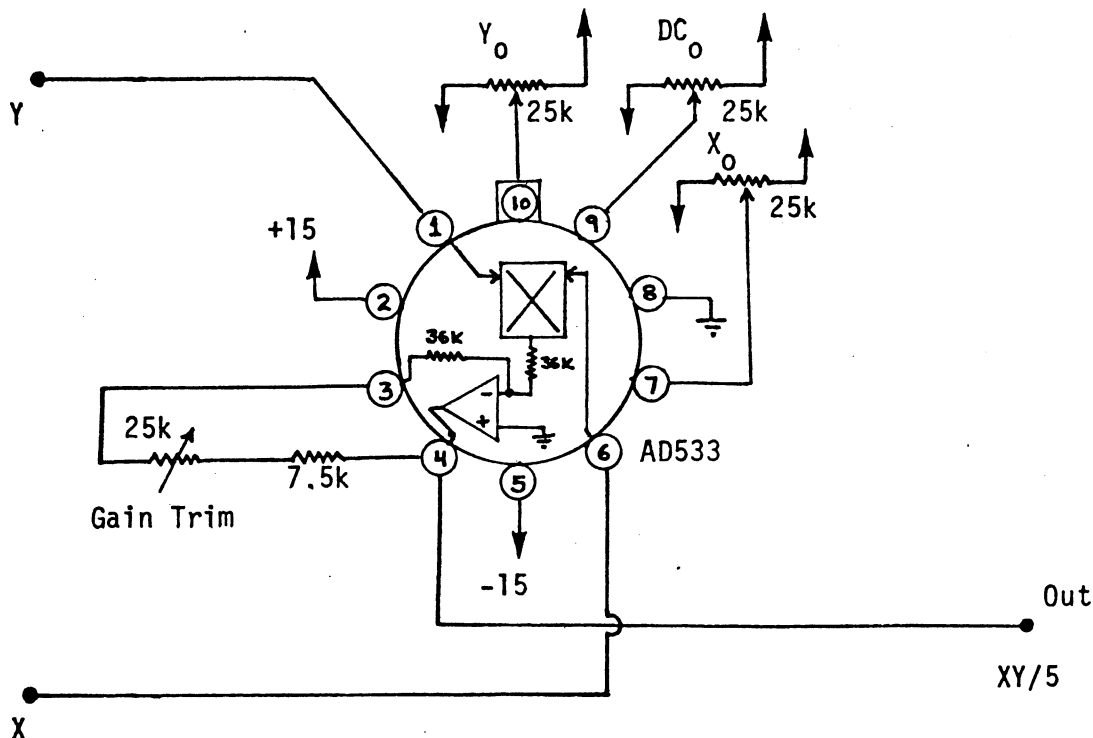
IC-3 forms a single stage all-pass phase shifter. If you like, you can think of this as a zero-bandwidth, infinite precision 90°PDN. Since we are feeding in only one frequency, we just adjust the RC time constant of the TP-2, R11, C5 string so that  $1/2\pi RC$  is 20 kHz. This provides a 90° phase shift. More information on this type of phase shifter can be found in AN-65 (to be published). In practice, TP-2 is adjusted to give this 90° phase shift. This 90° phase shift can be determined by a circular "Lissajous" figure on a scope (see AN-66, to be published).

Multipliers IC-4 and IC-5 form sum and difference frequencies (see Fig. 3) when the 20 kHz phases are multiplied with a sine wave from a VCO. The difference frequencies are in the range of 0-5 kHz. The sum frequencies are in the range of 35-45 kHz. Even though the sum frequencies are above the audible range, and will not become audible with later processing in this shifter, it is a good idea to filter them out. This is easy to do, and is accomplished with the two 2nd order Butterworth filters formed around IC-6 and IC-7. These have a cutoff frequency of about 5 kHz. By using this filtering, we know that the shifting signals will be cleaner and exactly what we think they are. We can also use these through zero signals for other purposes as well. Also, if we had left the sum frequencies in, they might cause interference in other modules further down the line, or cause problems in audio power amplifiers.

Fig. 7 shows the details of the multiplier circuitry. The multiplier is an Analog Devices type AD533 which costs under nine dollars in small quantities. These multipliers are much easier to use and perform much better than the type 595 devices we used in older designs. If you have used the older type, you will really appreciate these, and will see that the extra cost is justified. We have modified the circuit in the Analog Devices data sheet slightly so that the response is XY/5 instead of

XY/10. This fits our  $\pm 5$  volt levels better, results in better linearity, and allows us to avoid an attenuator on one input, as the gain control can be placed in the feedback loop of the internal op-amp. The trimming procedure is essentially the same as for any other multiplier, and is given in the figure. The same multiplier circuit is also used for the multipliers indicated in Fig. 11. Thus, you will need four circuits as shown in Fig. 7 for the complete shifter.

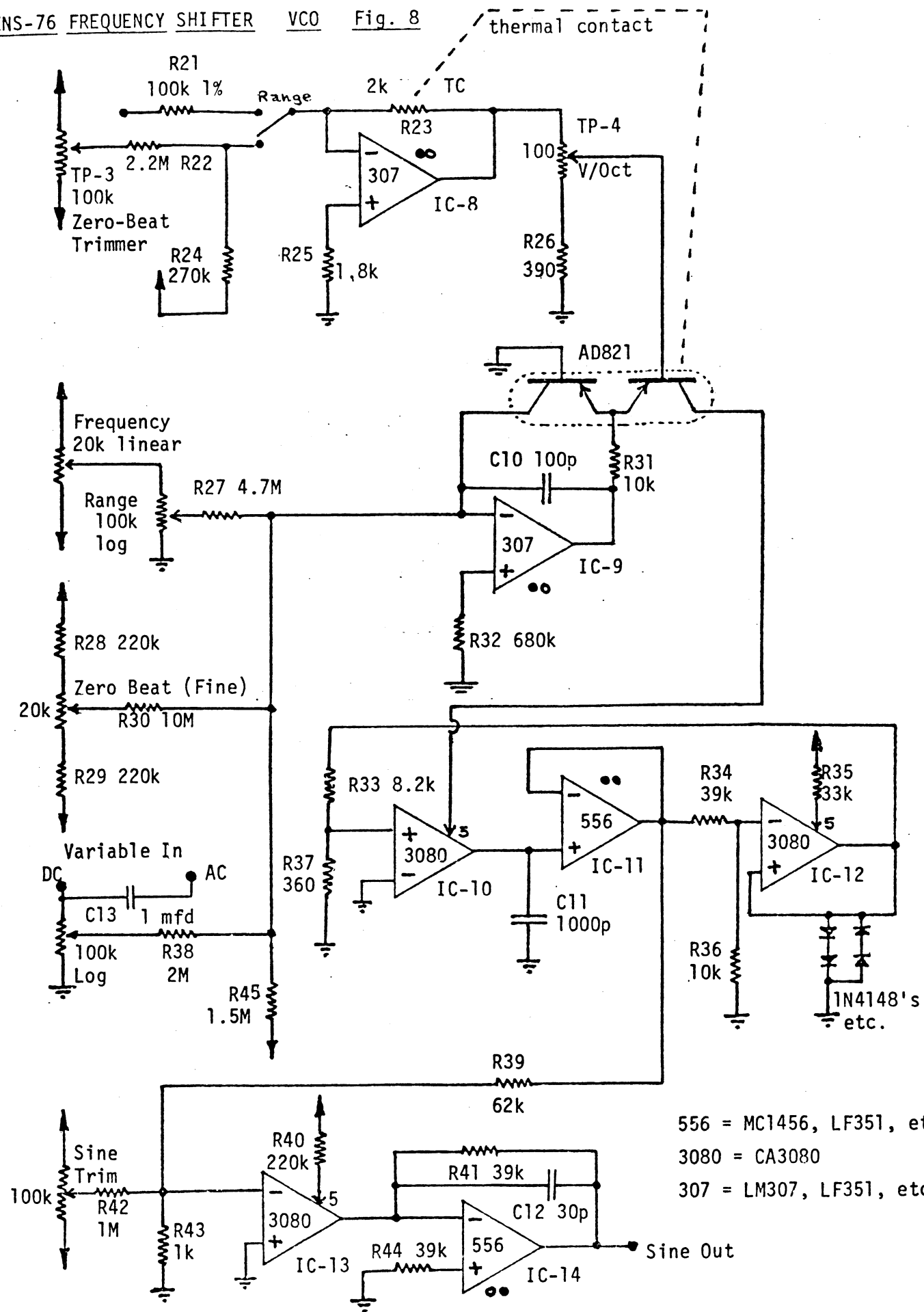
Fig. 7 Multiplier



#### TRIM PROCEDURE:

1. Adjust  $DC_0$  for zero volts out with X and Y grounded
2. Apply  $\pm 5$  volts 100 Hz to Y, ground X, and adjust  $X_0$  for minimum AC signal at the output.
3. Apply  $\pm 5$  volts 100 Hz to X, ground Y, and adjust  $Y_0$  for minimum AC signal at the output.
4. Ground X and Y. Readjust  $DC_0$  to zero volts if necessary.
5. Apply +5 volts to X and Y at same time. Adjust Gain Trim for +5 volts output.

The VCO circuit which we suggest is shown in Fig. 8. It is basically a modification of the ENS-76 VCO Option 2. With this circuit, we leave out the wave-shapers except for the sine, and have moved most of the frequency control to the linear control mode. With the range switch closed (down), the VCO is centered about 20 kHz. With the range switch up, the normal VCO range is available through a 100k 1 volt/oct. input. Bear in mind that this VCO is being used to beat against the fixed 20 kHz oscillator, and that 20 kHz (zero beat) corresponds to zero shift. A 15 kHz output of the VCO corresponds to -5 kHz shift while 25 kHz corresponds to a +5 kHz shift. Thus we are thinking of the VCO as working mainly in the region of 20 kHz. The combination "Range" and "Frequency" panel pots serve to set a range about 20 kHz and scan that



556 = MC1456, LF351, etc.

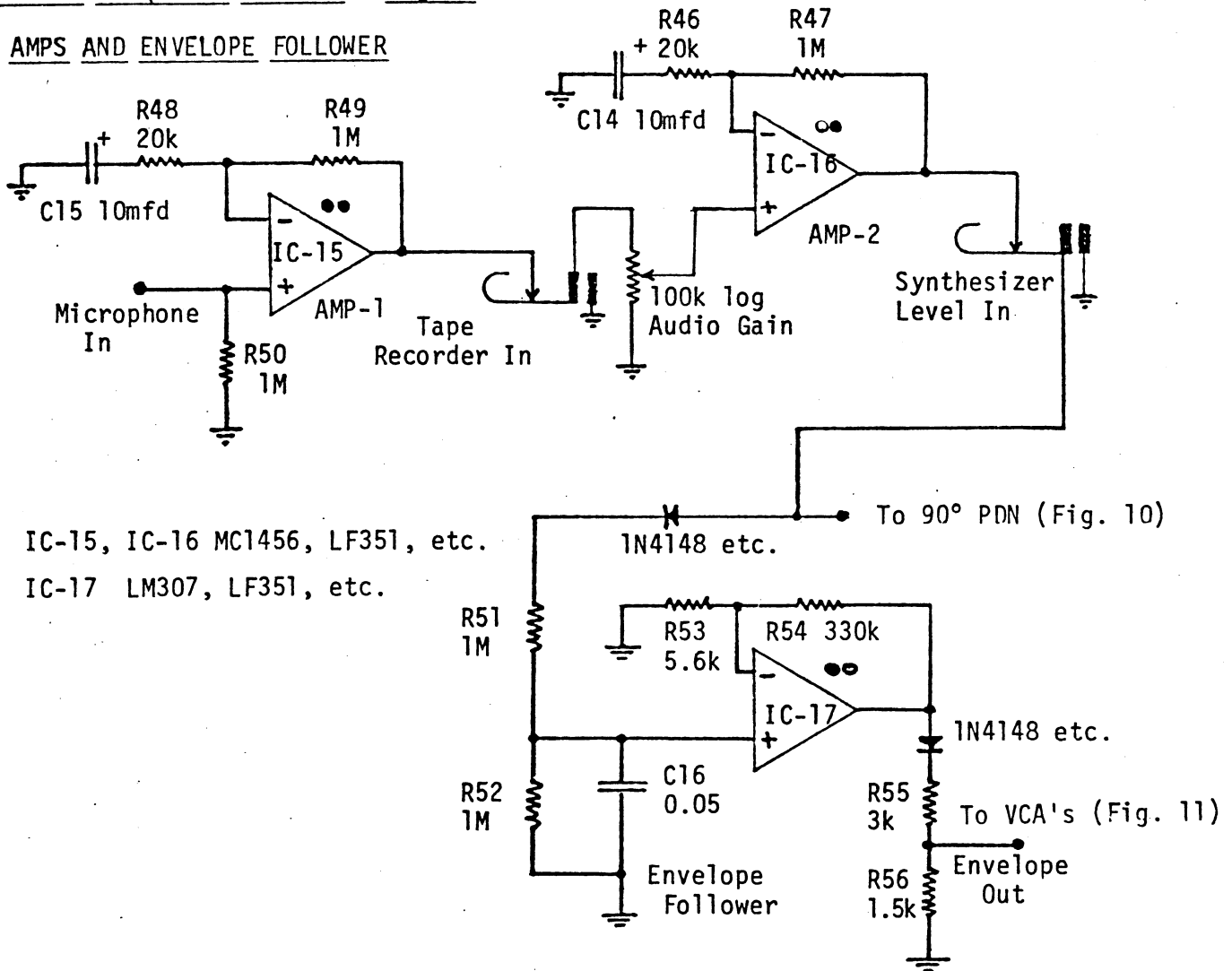
3080 = CA3080

307 = LM307, LF351, etc.

range respectively. The "Zero Beat" panel pot in this case serves to center the range or to fine tune as is necessary. Also, a variable input is available to receive modulating signals through either AC or DC coupling. More information on the VCO can be found in the ENS-76 VCO series in EN#75.

## ENS-76 FREQUENCY SHIFTER Fig. 9

### AMPS AND ENVELOPE FOLLOWER



IC-15, IC-16 MC1456, LF351, etc.

IC-17 LM307, LF351, etc.

Figure 9 shows the input amplifiers and the crude envelope follower. Each stage as shown has a gain of 50, and the combined gain is enough for most microphones. Note that switching jacks are used so that when a higher level signal is input, the amplifier(s) below are cut out. The middle jack is useful for tape recorder level signals while the synthesizer level jack accepts  $\pm 5$  volt signals directly. In any event, the audio signal is fed to the 90° PDN and to the crude envelope follower formed around IC-17. The 90° PDN is shown in Fig. 10 and will be discussed in a moment. The envelope follower shown is really more of a switch control because it has high gain (60). Thus, a quarter of a volt accumulated on C16 is enough to give a full +5 volts at the envelope out. This arrangement helps to fade in signals that are coming up but which are less than 0.85 volts (allowing 0.6 volts for the diode). Signals above this level are passed with unity gain since a +5 volt envelope is produced in all such cases. A 50 ms time constant was selected so that the envelope does not have sharp edges, and yet so that there is no noticeable delay once the signal appears at the input. Of course, the user is free to make any changes of time constant, gain, and any of the parameter can be made variable by panel pot. There may well be other combinations individuals will prefer.

The 90°PDN is shown in Fig. 10. The network shown is for a bandwidth of 15 Hz to 15 kHz with a phase difference of  $90^\circ \pm 0.2^\circ$  assuming infinite precision of the components. In practice, with 5% components and a little hand trimming of values we can expect to obtain an error of no more than 1-2° over the full range. As you probably know, it is not necessary to use the exact R and C values we show. You just have to keep the RC products constant. This is the same network we used in the earlier design [4] so we can use the same design data. Unlike the first design, we are using stages composed of single op-amps for the phase shifters [5] as there was no need to use the two amplifier design originally given. More information on these phase shift all-pass networks can be found in AN-65 (to be published). The data for this pair of networks is as follows:

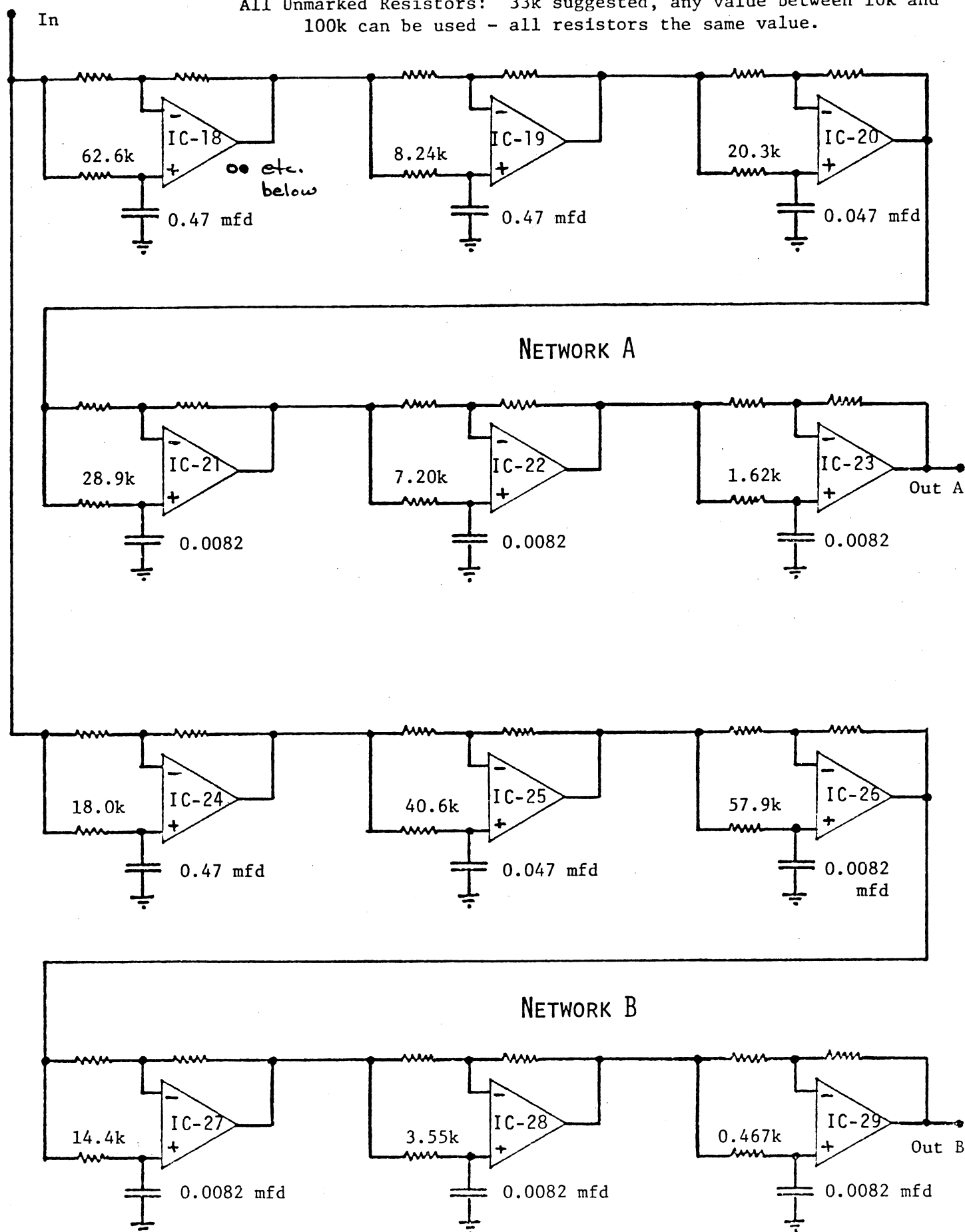
| <u>Network</u> | <u>Stage</u> | <u>90° Frequency</u> | <u>RC Product</u>     |
|----------------|--------------|----------------------|-----------------------|
| A              | 1            | 5.41 Hz              | $2.94 \times 10^{-2}$ |
|                | 2            | 41.1 Hz              | $3.87 \times 10^{-3}$ |
|                | 3            | 167 Hz               | $9.51 \times 10^{-4}$ |
|                | 4            | 671 Hz               | $2.37 \times 10^{-4}$ |
|                | 5            | 2694 Hz              | $5.91 \times 10^{-5}$ |
|                | 6            | 11977 Hz             | $1.33 \times 10^{-6}$ |
| B              | 1            | 18.8 Hz              | $8.47 \times 10^{-3}$ |
|                | 2            | 83.5 Hz              | $1.91 \times 10^{-3}$ |
|                | 3            | 355 Hz               | $4.75 \times 10^{-4}$ |
|                | 4            | 1344 Hz              | $1.18 \times 10^{-4}$ |
|                | 5            | 5472 Hz              | $2.91 \times 10^{-5}$ |
|                | 6            | 41552 Hz             | $3.83 \times 10^{-6}$ |

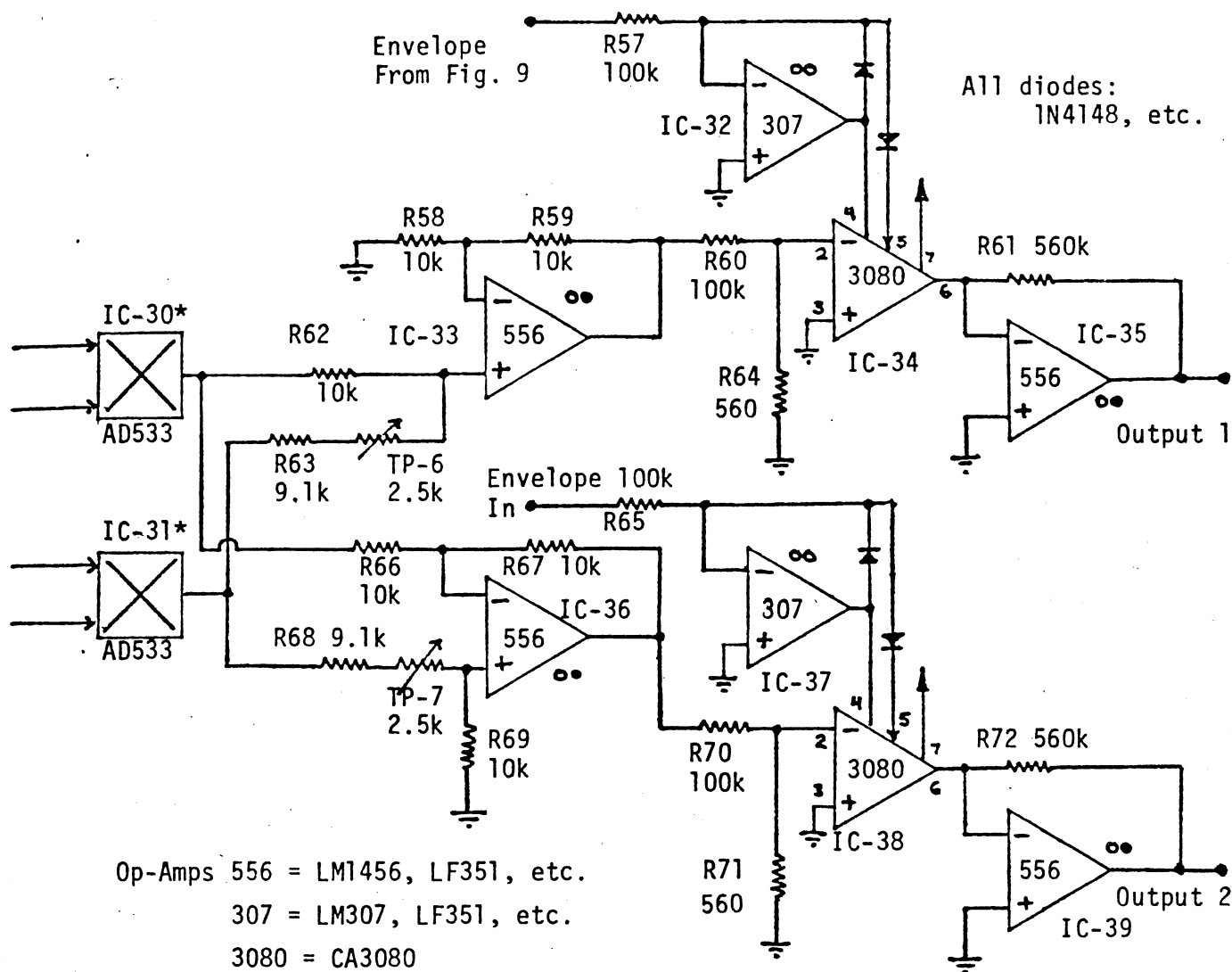
You may have all the required values of components available, but more likely, you would have to order some. You can just as easily use your own by making a simple calculation. Suppose for example you want to realize the third stage of network A (around IC-20) but don't have the 0.047 mfd capacitor. You do have a 0.02 mfd capacitor. The RC product for this stage is  $9.51 \times 10^{-4}$  so since 0.02 mfd is  $2 \times 10^{-8}$ , we get for R the value 47.55k. A logical starting point would then be a 47k 5% resistor. If necessary, you can trim this value up with a shunt or series resistor. It is a good idea to test each stage as it is constructed by the simple Lissajous figure method to see that it has a 90° frequency close to the value in the table. When all 12 sections are completed, you can obtain a circular Lissajous figure off the A and B outputs and this should be observed from 15 Hz to 15 kHz. This verification is enough to get the shifter going. Later if you think it necessary, you can tune up the stages individually as discussed in reference [4]. If you can get your hands on some 1% capacitors and have a digital ohmmeter, you can probably construct the network with sufficient accuracy without any tuning.

The final section of the frequency shifter is shown in Fig. 11. This section realizes the two multipliers, two summers, and two VCA's. The two multipliers IC-30 and IC-31 are configured as previously shown in Fig. 7. Each multiplier receives one signal from the 90°PDN of Fig. 10, and one signal from the quadrature oscillator of Fig. 6. Even if you get things mixed up, it usually comes out all right - you just get an upshift where you expected a downshift, and vice versa. The two summers are formed from IC-33 and IC-36. The VCA's are basically the same as the ENS-76 VCA. The first one is formed around IC-32, 34, and 35 while the second VCA is formed around IC-37, 38, and 39. The VCA's receive envelope control from Fig. 9, and both receive the same envelope. Note the unusual connection of pin 4 of IC-34 and IC-38.



All Op-Amps: MC1456, LF351, etc. (avoid use of 741 or 307 types)  
 All Unmarked Resistors: 33k suggested, any value between 10k and 100k can be used - all resistors the same value.





ENS-76 FREQUENCY SHIFTER, FINAL SECTION

Fig. 11

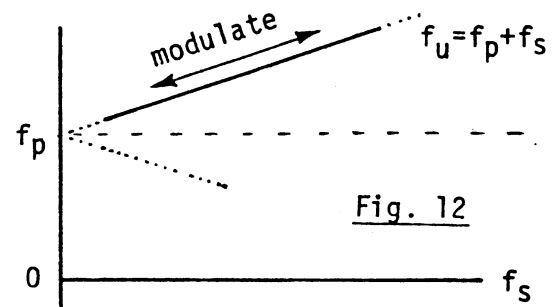
Once the shifter is completed and an initial check out is done (check the wiring, power supplies, and so on), you should test it with your own voice through a microphone. Since the effects achieved are at times unusual, it is essential that you have a clear reference as to what the input is, and you know your own voice as well as anything. Do not try to test with a signal from an oscillator as you will have to check frequencies all over the place to figure out what is going on. Note that there is a downshift and an upshift available at either output. Either output will also be capable of shifting your voice through zero and into another upshift. The two outputs are exactly the same except for the direction they are shifting. The reason for having two is that you may want to output two oppositely shifted channels (for a simulated stereo effect for example). If you checked out each section as you went along, it is fairly certain that the full scheme will work for you the first time. After you play with it a while, you can go back and tune things up a little more. Be sure you understand how the device is supposed to work so that you can understand how it is working for you.

This completes the circuit description. Before ending this section of the ENS-76 series, we want to say a few things about FM methods with the shifter, and discuss other applications.

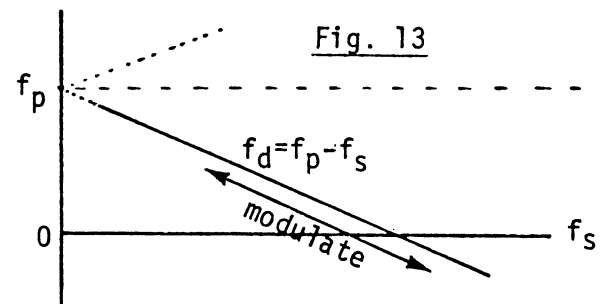
## THE FREQUENCY SHIFTER AS A FREQUENCY MODULATOR:

Since a frequency shifter is capable of altering a frequency, it is capable of frequency modulation. In most cases, the FM processes achieved can also be achieved with other devices such as VCO's and in some of these cases, the VCO is obviously simpler. In the case where the input is a live sound however, the frequency shifter offers processing not available with a simpler device. In the examination of the FM processes available from the frequency shifter, it is important that we keep in mind that the input is not just some frequency component  $f_p$ , but may be some complex live sound, otherwise we may dismiss a process as insignificant when actually it may be unique.

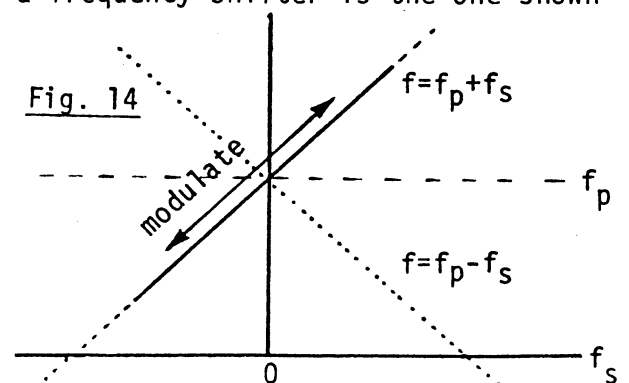
First we can consider a frequency shifter based around an ordinary quadrature oscillator which will not reach zero frequency or go negative. Let's look at the upshift first. The appropriate curve is shown in Fig.12. As the shifting frequency  $f_s$  varies, the output  $f_u$  varies accordingly such that  $f_u = f_p + f_s$ . Since  $f_s$  can be only positive,  $f_u$  is always above the input frequency  $f_p$ . The modulation process is linear FM only if  $f_s$  is a signal controlled in a linear manner. If  $f_s$  is produced by an exponential VCO, we won't be able to use dynamic depth FM any more than we can with the exponential VCO itself (unless of course we can tolerate the pitch shift). Thus the upshift case would seem to be of very little interest, since we need an FM process plus a frequency shifter to get another FM process which is much the same. However, as we mentioned in the first paragraph of this section, we must also consider live sounds as inputs. When we do we see that this process is useful. Suppose we want to modulate a human voice. We can upshift the voice and move the spectrum up and down, but we must always stay above the position of the original vocal spectrum. Thus you can upshift, but you can not down-shift at the same time with this system.



The second thing to look at is the downshift output. This is much the same as the upshift up to the point where the frequency reaches zero going down. The appropriate shifting curve is shown in Fig.13. Here we can shift down through zero to negative frequencies, even though the frequency of the quadrature oscillator does not even reach zero. This is something that is of interest even for inputs as simple as a sine wave from an electronic oscillator. Of course, live sounds are even more interesting. The frequency shifter thus offers a means of achieving through-zero FM alternative to the design of phase reversing VCO's (see the ENS-76 VCO Option 4 in EN#75).

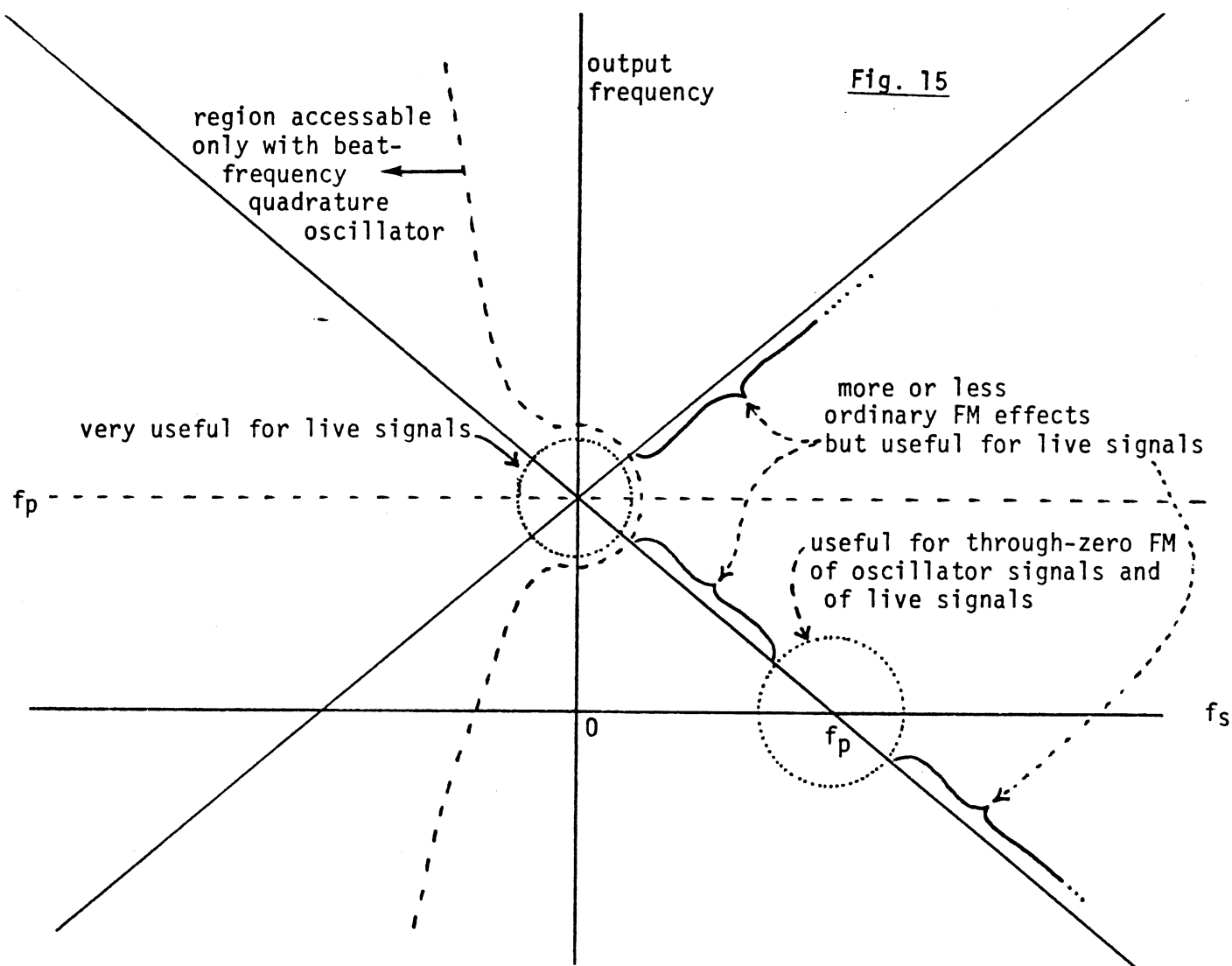


A third FM process that involves the use of a frequency shifter is the one shown in Fig.14. Here we need a quadrature oscillator that is capable of passing through zero frequency. Thus we must assume that the beat-frequency method is being used. This region is not accessible with the ordinary quadrature oscillator. Since the beat-frequency quadrature oscillator requires a more complicated circuit, we have to ask whether it is worth it or not. We can see that the case where an oscillator signal is processed is relatively uninteresting since this is really just ordinary FM. The



corresponding case where a live signal is processed is quite interesting however. This is because we can shift the live signal through zero shift (not through zero frequency). A zero shift point is interesting if we have a clearly referenced unshifted signal. The sound of our own voice is a good example - we know what it should sound like. If the voice is shifted up from its initial spectrum, back down to the starting point, then down below, back up to the starting point again, and so on, we will be aware of a dramatic effect occurring. On the other hand, if the input signal is an oscillator signal, we only are aware of FM effects. To put it another way: if a synthesized signal is processed, the processing step will seem to be just another synthesis step.

By way of summary of the useful FM processes with a frequency shifter, Fig.15 is presented below. Study of this diagram and a rereading of this section should make it clear which areas are useful.



#### ADDITIONAL APPLICATIONS:

In addition to the frequency modulation of various live and synthesized sounds, the frequency shifter has a number of other important applications. Many of these have been discussed in this newsletter, and others were described by Bode and Moog [2]. Almost certainly more applications would be found if frequency shifters were found in more studios. Here we will discuss some of the other uses of the component parts of the frequency shifter module. After reading this, you may want to include some extra output jacks on your unit.

First, it is fairly obvious that there are a number of units that can be used directly for other purposes. First, there is the microphone amplifier which can be used to amplify vocal sounds to a synthesizer level for processing by synthesizer modules. Secondly, there is the VCO which can be used as a regular VCO, or at least as an extra low-frequency sine wave oscillator. There are also four excellent balanced "ring" modulators in the system (the multipliers) and these may be made available by using switching type jacks on the panel of the unit.

It is important also to realize that the quadrature oscillator itself provides a frequency modulation signal that passes through zero frequency. Thus we can use the beat-frequency quadrature oscillator itself to produce through-zero FM. Thus there are two ways of providing through-zero FM with the shifter unit. This is a net plus for the unit although it makes the production of through-zero FM through shifting less important. [It remains important with the ordinary quadrature oscillator.] Probably there are some interesting effects that can be achieved by using two effects together.

There are also some interesting effects that can be achieved in video synthesis with the unit. Quadrature oscillators have long been used in video synthesis. Another point is that the 90°PDN is useful for video synthesis. If you form a Lissajous figure off the ends of the 90°PDN you will soon see that there are many interesting patterns that can be formed (complex circular swirls) by inputting complex waveforms and live sounds.

#### REFERENCES:

- [1] Harald Bode, "Solid State Frequency Spectrum Shifter" AES Preprint 395, Oct. 1965
- [2] Harald Bode and Robert Moog, "A High-Accuracy Frequency Shifter for Professional Audio Applications," J. Aud. Eng. Soc., Vol. 20, No. 6, July/August 1972, pp. 453-458
- [3] Harald Bode, "Apparatus for Producing Special Audio Effects Utilizing Phase Shift Techniques," US Patent 3,800,088, March 26, 1974 (filed Aug. 28, 1972)
- [4] B. Hutchins, Musical Engineer's Handbook, (1975) Chapter 6a
- [5] B. Hutchins, "Phase Shift Networks and Applications" EN#59 (2) Nov. 1975
- [6] Reference [2] page 455
- [7] Data and Applications Literature on AD533 multipliers and related products. Analog Devices, PO Box 280, Norwood, MA 02062
- [8] Norlin Music (Moog); Bode Sound Company, N. Tonawanda, NY 14120; 360 Systems Los Angeles, CA

\* \* \* \* \*



## SAMPLE AND HOLD UNITS

THE SAMPLE-AND-HOLD UNITS: The two sample-and-hold options here are basically the same circuit, and the second has a few additional features. Either circuit is suitable for relatively slow sampling (10 Hz or slower) and is useful for sampling white noise, for breaking slow control waveforms into steps, and for similar purposes. The second option has a feature for single step triggering, and a slew limiting feature that can be used to add correlation to sequences of sampled random noise.

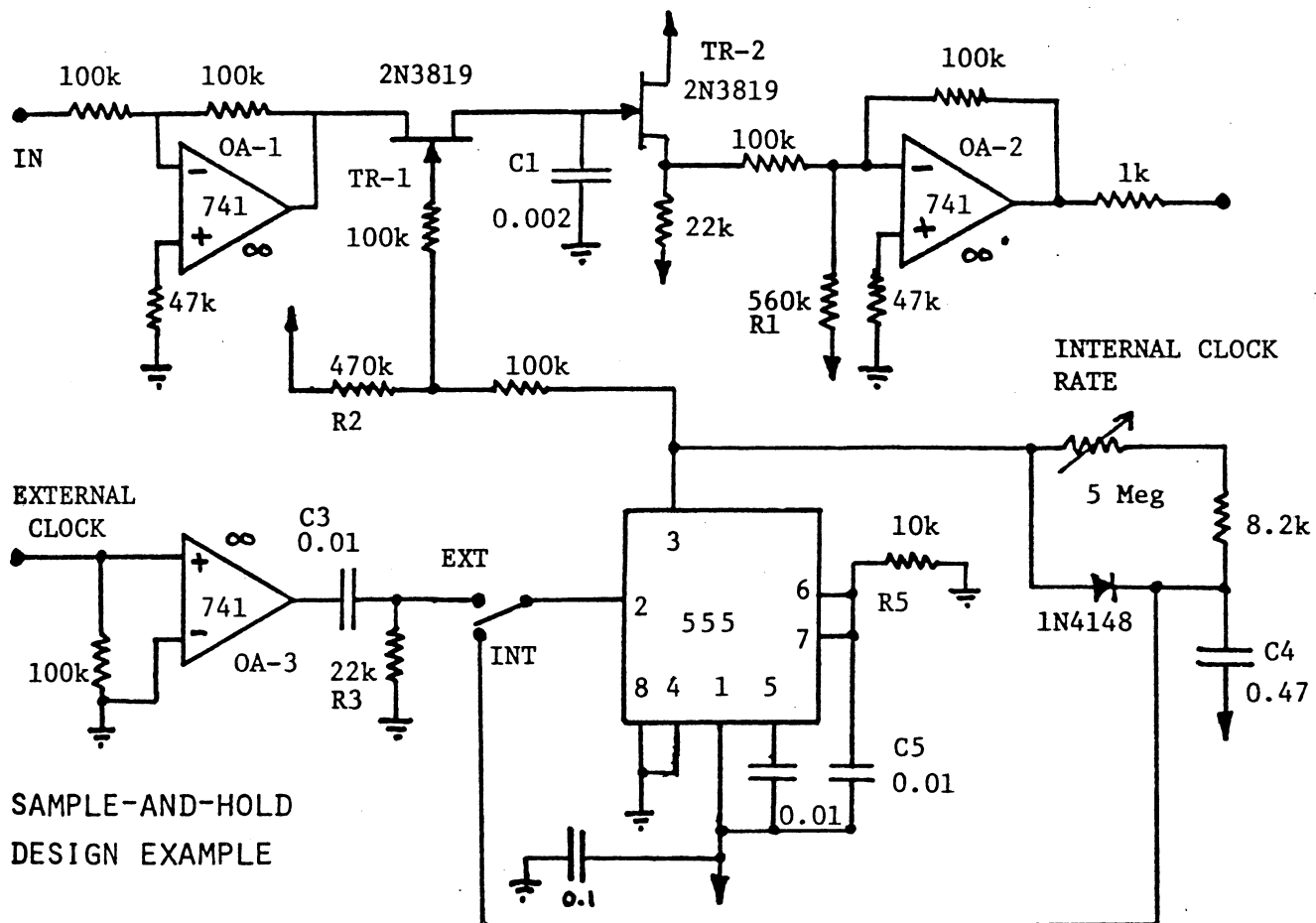
| <u>Preferred<br/>Circuit<br/>Option</u> | <u>Original<br/>Source</u> | <u>Original<br/>Designer</u> | <u>Estimated<br/>Parts<br/>Cost</u> | <u>Notes</u>                            |
|-----------------------------------------|----------------------------|------------------------------|-------------------------------------|-----------------------------------------|
| S&H-1                                   | MEH 5g                     | Hutchins                     | \$4                                 | Int. & Ext. Trigger                     |
| S&H-2                                   | EN#61                      | Hutchins                     | \$5                                 | Slew-limiting and<br>single step added. |



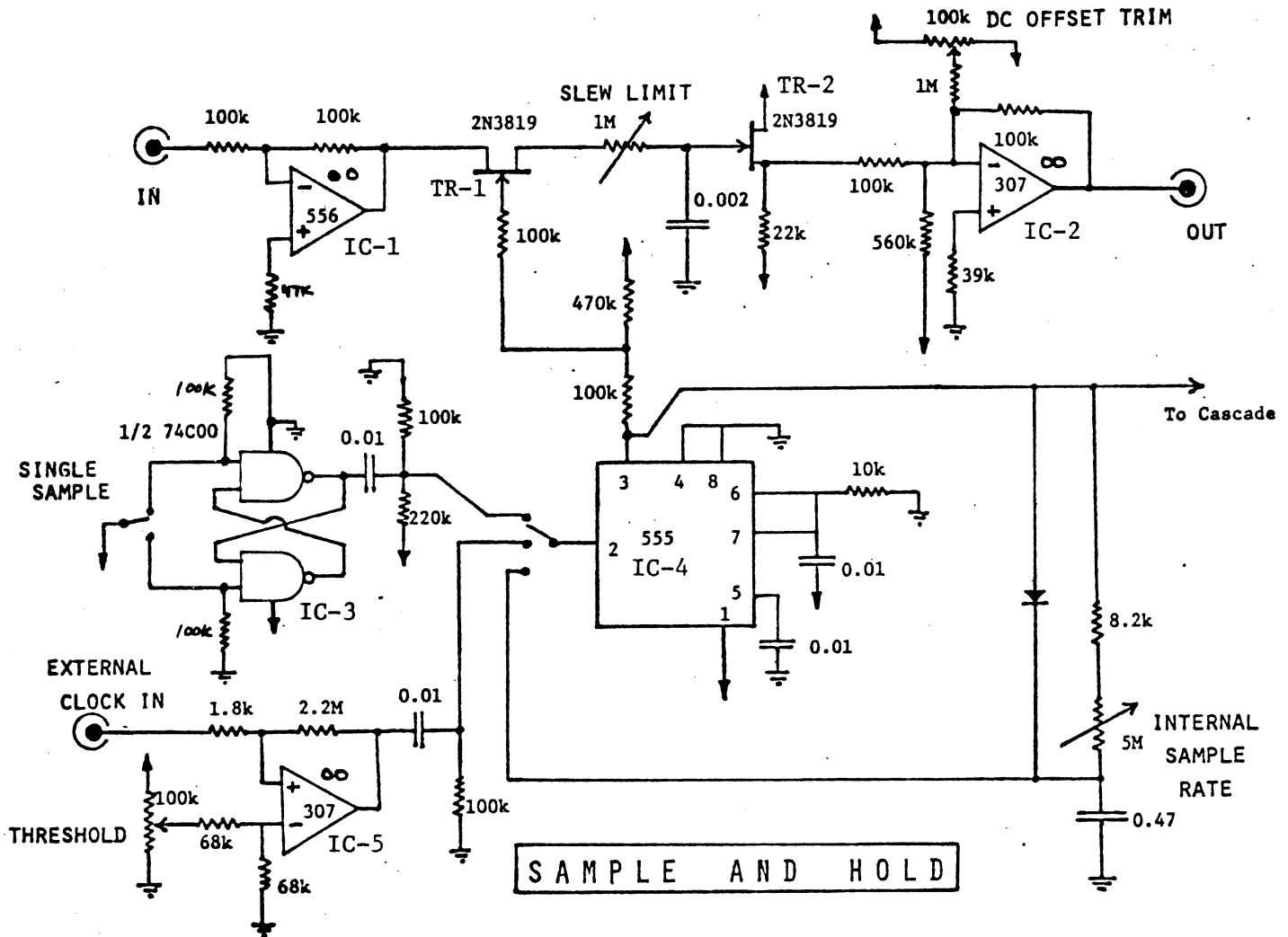


In the design example, TR-1 serves as the sampling switch. Since it is an N-FET, it is held off by a negative voltage applied to the gate (negative voltages repel the negative electrons which are the charge carriers). The FET switch is turned on when the gate voltage is allowed to float up to the input voltage from OA-1. C1 serves as the holding capacitor, and the voltage on C1 is buffered by source follower TR-2. This source follower is necessary to prevent excessive drain of current from the capacitor. However, the FET buffer causes a substantial offset voltage which is excessive in this case. [In many cases, the FET is inside a feedback loop and the offset is not important. In other cases, the offset is nulled out later on.] It is convenient to correct for this offset using the inverting summer (OA-2) and the op-amp is also useful for lowering the output impedance of the device as a whole. To compensate for the inversion, the input buffer (OA-1) is made an inverter as well. The input impedance is thus the standard 100k. It can be seen therefore that the actual voltage is stored as the negative of the input. Resistor R1 serves to compensate for the offset of the source follower. If a different FET is used, this value may have to be changed some.

The heart of the timing section is the type 555 timer. Note that this is essentially the standard monostable configuration for a 100  $\mu$ sec. pulse out ( $C5 \cdot R5$ ). The 555 is powered between ground and -15 instead of the usual +15 to ground. This makes no difference to the 555, but the pin 3 out is thus normally held down around -15, and rises to give a 100  $\mu$ sec. pulse rising toward zero. The resistor R2 serves to pull this output up (as seen by the gate of TR-1) and this allows the FET switch to be clocked with a gate voltage in the proper range.



The second S&H Unit is similar to the first except it has some additional features. First, it has a slew limiting option added which makes it possible to limit the maximum swing between successive samples. If the change is too much, you just get part of the change, not all of it. This tends to correlate samples. The second feature is the single sample feature. A third feature is a variable threshold on the trigger.



IC-1 serves as an input inverter/buffer. TR-1 serves as the sampling switch while TR-2 forms a source-follower buffer for the voltage held on the 0.002 mfd sampling capacitor. IC-2 serves to invert the held voltage and adjust the DC level of the output. IC-4 is a 555 timer which operates in either an astable mode or a monostable mode. In either case it provides pulses (from a -15 level to ground and back to -15) of about 100 microseconds that serve to trigger the FET switch TR-1. This pulse appears at pin 3 of the 555 and the attached resistor network gives the proper voltage at the gate of TR-1 to turn it off (pin 3 at -15) or on (pin 3 at ground). IC-4 can be triggered by IC-3 which forms a bounceless pushbutton for single manual samples. IC-4 can also be clocked externally from IC-5. Note that a variable threshold can be set for IC-5. This is very useful for triggering from a noise source so that the sample times are random.

## SLEWING CIRCUITS

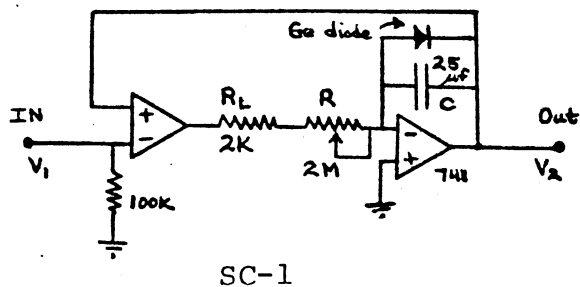
THE SLEWING CIRCUIT OPTIONS: The first three of the slewing circuits are manually controlled devices, while the fourth is a voltage-controlled slew limiter. These devices are used to limit the rate of change between two voltages. These are variously called portamento circuits, slew limiters, and lag processors. The slewing devices are mainly used to process control voltages.

| <u>Preferred<br/>Circuit<br/>Option</u> | <u>Original<br/>Source</u> | <u>Original<br/>Designer</u> | <u>Estimated<br/>Parts<br/>Cost</u> | <u>Notes</u>                       |
|-----------------------------------------|----------------------------|------------------------------|-------------------------------------|------------------------------------|
| SC-1                                    | EN#42                      | Titchener                    | \$2                                 | Linear                             |
| SC-2                                    | EN#42                      | Mikulic                      | \$2                                 | Linear                             |
| SC-3                                    | EN#42                      | Mikulic                      | \$2                                 | Exponential                        |
| SC-4                                    | EN#59                      | Titchener                    | \$4                                 | Lin/Exp.<br>Voltage-<br>Controlled |



4a. READER'S EQUIPMENT: The following slew limiting ideas were provided by Paul Titchener.

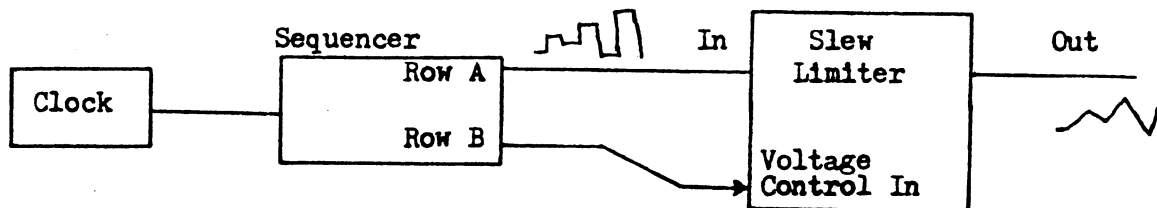
The circuit at the right is a slew limiter or portamento that glides between voltages in a straight line fashion instead of the usual exponential slope. It can be made voltage controlled with the addition of a 595 multiplier or one of the CA3080 Voltage Controlled Resistors presented by Gordon Wilcox in EN#40 (replacing R).



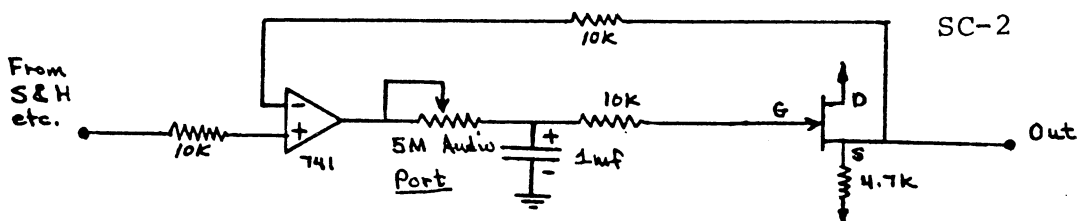
The circuit uses an integrator and a comparator. When  $V_1 \neq V_2$  the comparator switches the integrator to charge C until  $V_1 = V_2$  at which point the comparator oscillates at a high frequency keeping  $V_1 = V_2$ . Slew Rate =  $dV_C/dt \approx 14/(C(R+R_L))$ .

When used with a keyboard, the portamento effect is noticeably different from the "usual" type portamento, although not easily used due to unequal time intervals between note interval steps.

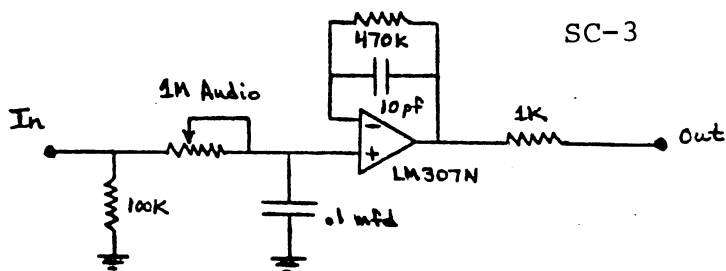
An interesting use of the voltage controlled version is to feed the output of a row of a sequencer through the limiter, and use another row to control the slew rate. This makes a "point-by-point" waveform generator with the slope of the sections controlled through the individual sequencer channels adjusting the slew rate.



A second example of the same basic circuit is provided by Terry Mikulic's linear portamento circuit below. Terry's circuit does not use the diode in the integrator, and will swing to full  $\pm$  supply (of course you can do the same thing by pulling the diode out of Paul's circuit too).

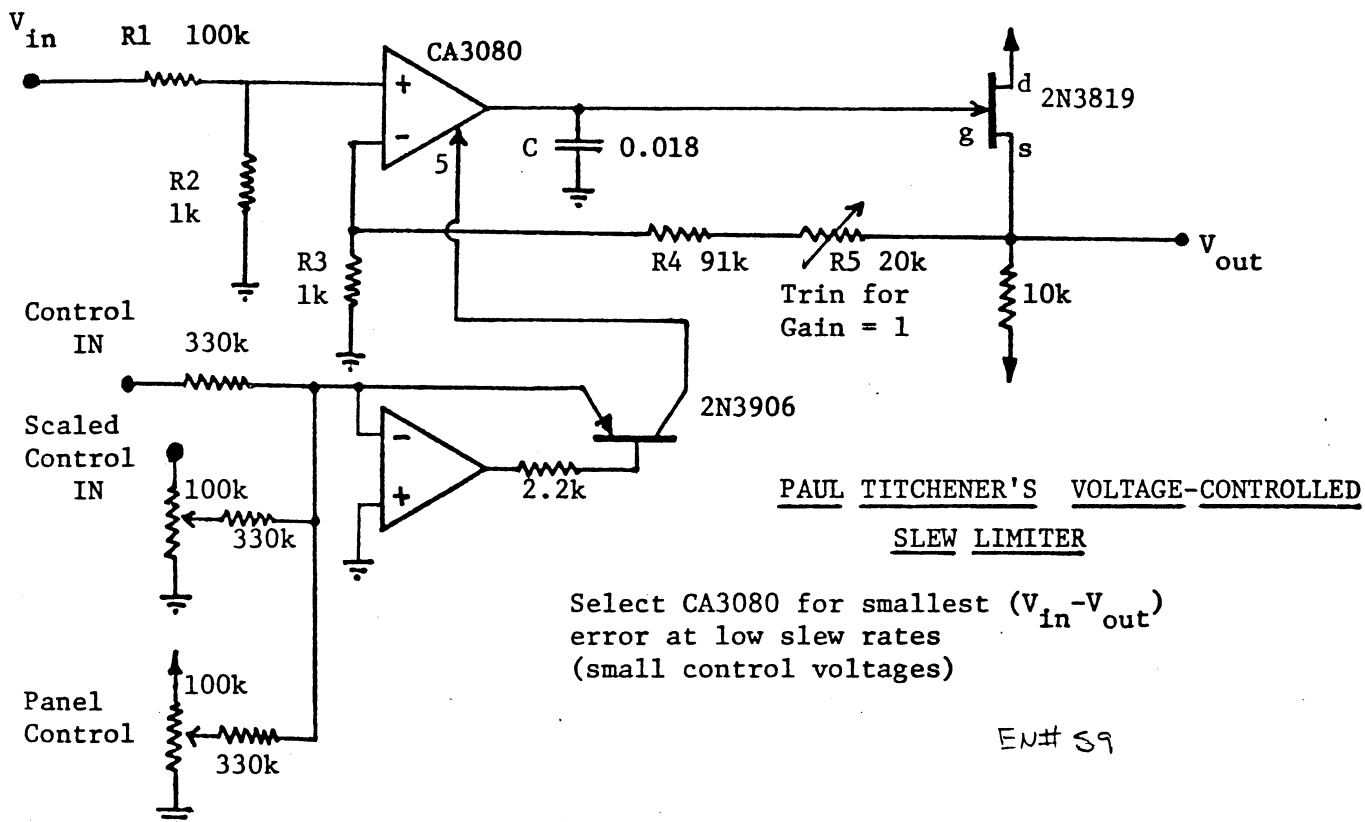
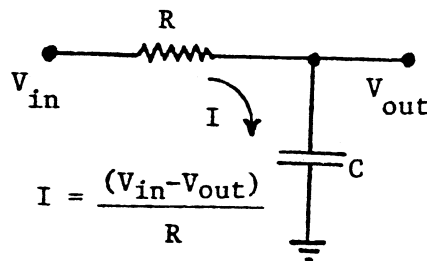


A related circuit is the "Lag Processor" shown below, also from Terry Mikulic. Terry indicates that the lag processor is actually a portamento control, can be used as a low-pass filter, but is most commonly used to modify rapidly changing control voltages. Note that portamento in this case is exponential.



EN#42

The voltage-controlled slew limiter circuit shown on the top of the next page was submitted by Paul Titchener, 4021 Rosehill Ave., Cincinnati, OH 45229. The circuit provides standard "exponential rise" portamento as shown. Note that the standard form of an exponential portamento circuit (as shown at the right) has a current into the capacitor that is proportional to  $(V_{in}-V_{out})$ . In Paul's circuit the resistor  $R$  is replaced by the CA3080 acting as a voltage controlled resistor. In order for the current out of the CA3080 to be proportional to  $V_{in}-V_{out}$ , it is necessary that the input differential voltage be limited to the linear region for a two transistor differential amplifier (which is what the input of the CA3080 is). This means a limit of about 40 mv or so, and this is the reason for the 1k attenuating resistors  $R_2$  and  $R_3$ .



Paul mentions the following simple modification that changes the circuit so that it produces linear rather than exponential ramps:

$$R_1=R_4= 2.2k$$

$$R_2=R_3= \infty$$

$$R_5=0$$

Note that this removes all attenuation from the inputs of the CA3080. This means that the inputs will in general be saturated (differential voltage input exceeds about 40 mv) at all times except when the slew has been completed. This is because the saturated input means that the control current to the 3080 passes to the output. This means that the capacitor is being charged by a constant current that is dependent on the control voltage, but independent of the input voltage, and this means a linear ramp. Note however that in the linear mode the actual differential input voltage of the 3080 should not exceed  $\pm 5$  volts, as this is the maximum value allowed for the 3080.

Paul further notes that at low slew rates (small control voltage), there is an error voltage  $(V_{in}-V_{out})$  of about 0.165 volts, apparently due to the decreasing gain in the loop. (Also possibly due to the varying differential input resistance). This error is audible when for a constant  $V_{in}$  the control voltage is changed rapidly from a small to a large value. This error could be compensated for by summing an inverted and appropriately scaled component of the control voltage to the output.

## NOISE SOURCES

THE NOISE SOURCES: The options for this collection with regard to noise sources begins with the ENS-76 circuits. The ENS-76 Option 1 is a standard white noise source formed from a back biased transistor. A number of accessory circuits are also given, and these provide pink noise, low-frequency random noise, and a random vibrato source. Also in the ENS-76 series is a digital "Pseudo-Noise" source which is given as the second option to this collection. The output of this circuit is a white noise which can also be processed with the accessory circuits of the first circuit. Also in this collection is a somewhat simpler white noise source from the ENS-73 series, and a noise source based on a noise generator IC.

| <u>Preferred<br/>Circuit<br/>Option</u> | <u>Original<br/>Source</u> | <u>Original<br/>Designer</u> | <u>Estimated<br/>Parts<br/>Cost</u> | <u>Notes</u>             |
|-----------------------------------------|----------------------------|------------------------------|-------------------------------------|--------------------------|
| NS-1                                    | EN#76                      | Hutchins/Mikulic             | \$2                                 | ENS-76 Series            |
| NS-2                                    | EN#76                      | Hutchins                     | \$6                                 | ENS-76 Series<br>Digital |
| NS-3                                    | EN#30                      | Rossum                       | \$2                                 | ENS-73 series            |
| NS-4                                    | EN#64                      | Hutchins*                    |                                     | Noise IC                 |
| NS-5                                    | AN-143                     | Hutchins                     |                                     | Overview                 |

\*straightforward design from application notes





NS-1 and NS-2 are part of this entire reprint from EN#76  
THE ENS-76 HOME-BUILT SYNTHESIZER SYSTEM - PART 8, RANDOM SOURCES:

-by Bernie Hutchins, ELECTRONOTES

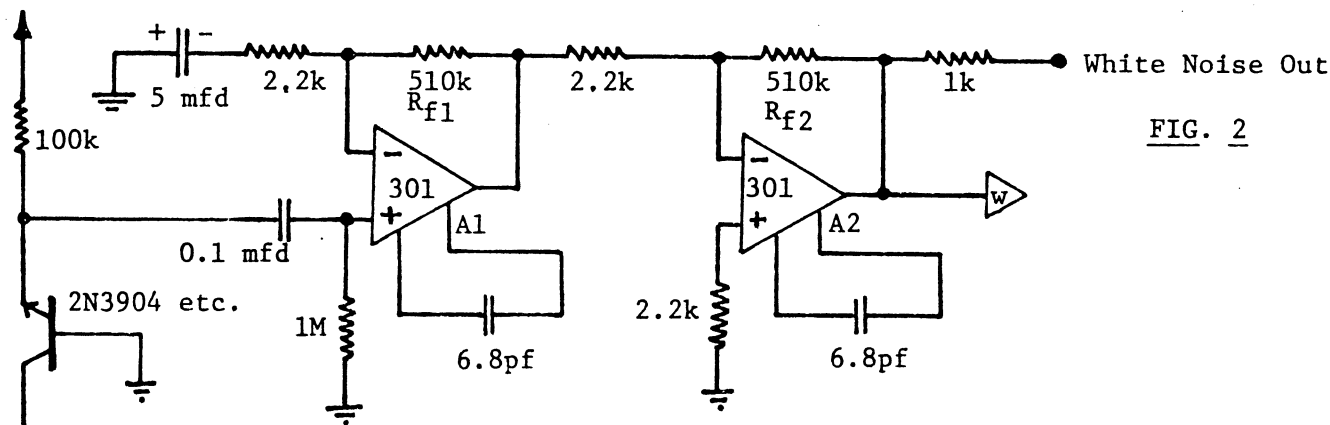
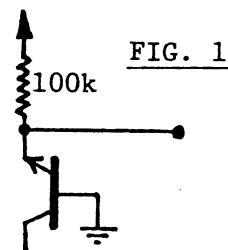
Most synthesizers have one or more "Noise Sources" which provide outputs for "White Noise" and often for "Pink Noise" as well. White noise has a flat spectral density, while pink noise has a spectral density that rolls off at 3db/octave, thus providing equal spectral energy in all bandwidth of the same ratio. Both types have a general character of "static noise" and are very similar to the interstation noise that is heard on an FM receiver. These noise sources can be used to provide basic textural material for sound synthesis, can be used for the synthesis of noise-like sounds (e.g., thunder), or can be converted to control signals with devices like filters and sample-and-hold units.

There are two common approaches to the generation of random signals. First, you can use true random noise generated in an electrical device as a result of some physical process that is random. A back-biased transistor junction is a good source of such noise if it is properly amplified. The second approach is the "Pseudo Random" (PR) approach which employs digital feedback shift registers to generate a long series of digital states in a sequence that "seems" random to non-specialized electrical instruments and to the subjective human ear.

In our approach, we shall offer several random units. A simple white noise source based on the back biased transistor will be presented for simplicity and economy. We shall also offer some PR devices with additional features. Finally, we want to look at devices for generating sounds that have both random (or PR) aspects as well as periodic aspects to them. These latter devices are new and unique devices which we have not discussed before.

### ENS-76 RANDOM SOURCE - OPTION 1

The simplest form of the back-biased transistor junction noise source is shown in Fig. 1. Note that it consists of only a single transistor and a current limiting resistor. The noise voltage generated at the junction is not large enough for synthesizer use (although it has a very wide bandwidth), so for an actual source, we need quite a bit of amplification so that we get something like a five volt level. It is difficult with these sources to set a level since the signal is random, but the user should have little trouble getting a satisfactory signal. The circuit for the White Noise Source is shown in Fig. 2



The circuit of Fig. 2 is virtually unchanged from previous designs. A1 and A2 form a two stage amplifier. The gain of these amplifiers is adjusted so that the noise level at the output is sufficient. This depends on the individual transistor

selected for T1, so you may have to do some adjusting of the feedback resistors  $R_{f1}$  and  $R_{f2}$ , upward for more gain, and downward for less gain. If desired, trim pots can be used to make this easier. We have selected type 301 amplifiers here (or substitute type 748) because these are easily compensated for higher gain giving a larger bandwidth. Type 741, 307, should not be used here. Type 556 is acceptable, but the 301 or 748 will probably have a "crisper" sound.

It is typical to do some processing of the white noise output to make other types of random signals available. These are driven off amplifier A2 (from the little triangle with the "w" inside). Below we show three processors. Fig. 3 shows a pink noise filter and a low-frequency random signal processing filter. There are from a design submitted by Terry Mikulic and first presented in EN#34. A graph of the frequency response of the pink noise filter can be found in EN#64 (14), and the report on noise generators given in that issue may be useful reading at this point. Note that the random signal is driven off the pink noise output. It can also be driven directly off the white noise, although some gain adjustments may be needed in this case. Be a little careful adjusting the level of the L.F. Random Signal as it has a relatively small bandwidth (below 10 Hz) and should be observed for 10 seconds or more before you reach any conclusions about the signal levels you might expect. If you observe for too short a period, you may set the gain too high. The third processing unit is the 7 Hz bandpass filter shown in Fig. 4. This is a high-Q bandpass filter which is excited by the white noise. Its output depends on the exact setting of the Q control, but tends toward a 7 Hz sinusoidal as Q is increased. We put this in to provide a somewhat random vibrato signal. With the Q turned to the max position, the circuit oscillates providing a steady state vibrato signal. The design of the filter is discussed in EN#73 (14).

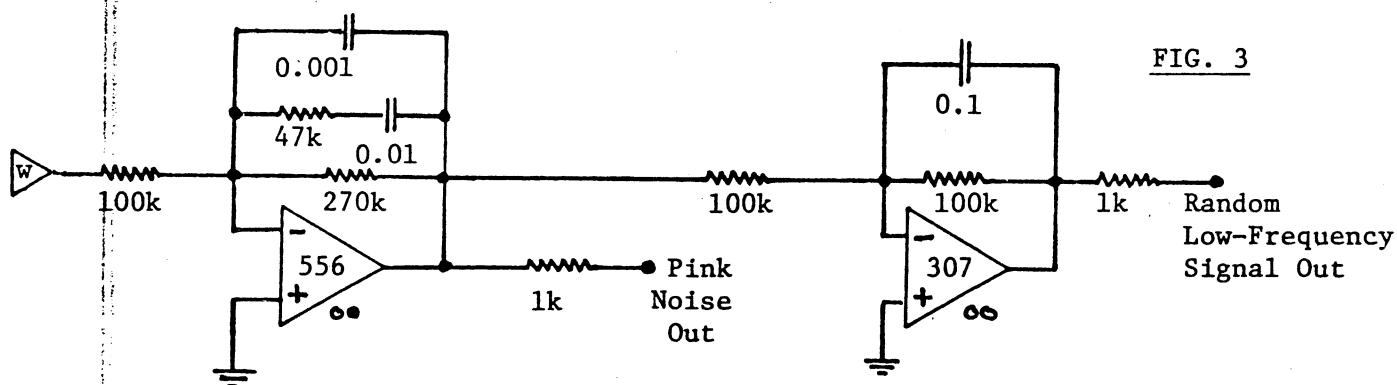
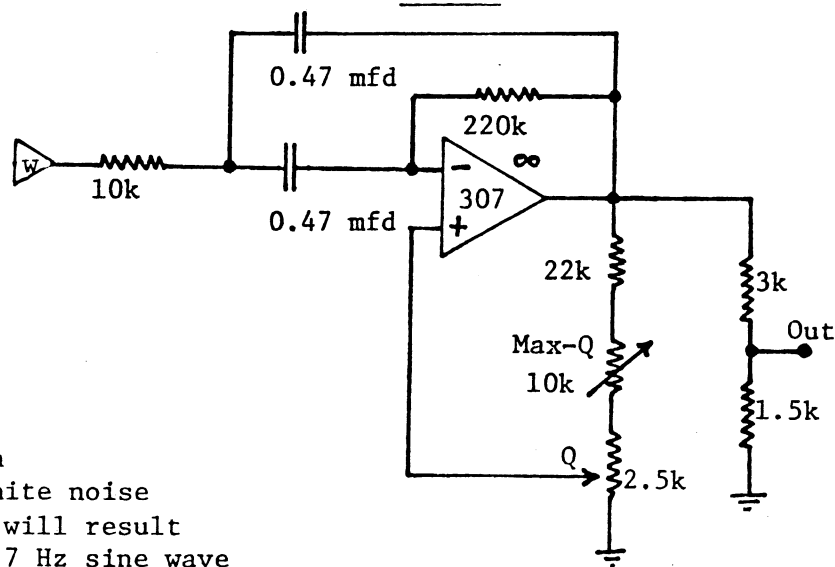


FIG. 4 Random Vibrato Source



The bandpass circuit is adjusted by first grounding the input. Then the Q control is set all the way up, and the Max-Q trimmer is then set so that the circuit oscillates with only a small amount of clipping. It is possible to adjust the Max-Q control so that a reasonably pure sine wave output is achieved, but the level will not be stable. Thus, a small amount of clipping should be used so that oscillation is steady. Now connect up the white noise source. Decreasing the Q control will result in a signal that is essentially a 7 Hz sine wave with randomly varying amplitude, and as the Q is

further decreased, the output becomes more and more random in frequency as well as in amplitude. Also, the amplitude will drop as the Q is decreased, but this can usually be compensated for in normal use by changing the input gain of the circuit being modulated by the vibrato signal.

## ENS-76 RANDOM SOURCE - OPTION 2

In Option 2, we will be replacing the white noise circuit of Fig. 2 with a Pseudo-Random (PR) shift register. This is more complex and more expensive than the simple back-biased transistor. So why do it? There are several reasons for doing this (or for considering doing this - as the case may be). The most important reason is that the resulting outputs will be repeatable, although seemingly random. If something of particular musical use does occur in the output, we will at least have the potential of recovering this and doing it again. This is done by resetting the generator or letting it go through a full cycle. The reader unfamiliar with PR generators should consult a text on feedback shift registers, and/or read the discussion of noise generators in EN#64. A second reason for considering a PR generator is that it is much easier to set up and maintain (from unit to unit) signal levels in the generators since the noise output does not depend on the individual diodes (back-biased transistors) used. A final reason for choosing the PR generator is that a lot of additional units can be driven from the PR generator. We will be considering these units in final options. They will include units for different statistical distributions of amplitudes, recycle units for random selection of timbre and for randomly changing timbres, and units for periodic changes of random timbre.

The basic white noise generator of Option 2 is shown in Fig. 5. The circuit uses a 24 stage shift register with Modulo-2 feedback according to the setup required for a maximal length sequence. [See Chapter 5h of MEH]. The total length of such a sequence is  $2^n - 1$  where n is the number of stages - hence this 24 stage shift register produces a PR sequence of length 16,777,215. At the 36 kHz rate of clocking shown, the sequence will cycle after a period of nearly 8 minutes. This is more than long enough for most purposes. We selected this length because it was exactly the length you get using three type 74164 chips, and two chips (shift register length 16) is not long enough. The start up circuitry shown is needed to prevent the shift register from starting up in the "All Zero" state, from which it can never emerge without special resetting. The output of the Exclusive-OR gates, or any other stage of the shift register, is the PR Binary Sequence that we will be using. The output sequence could in fact be used directly as a white noise source. However, while the spectrum is right for white noise (i.e., flat over the audio range), the statistical distribution of amplitude levels consists of only two possible values, unlike the back-biased junction source that is approximately Gaussian (see EN#62). It is well known however that low-pass filtering of the PR Binary Sequence will result in an amplitude distribution that has continuous levels, and approaches Gaussian. The 1k and 0.01 mfd R-C filter gives a cutoff of about 16 kHz, which is enough in this case, although not really enough to assure the best approach to a Gaussian distribution. If the best Gaussian distribution is desired, the 0.01 mfd capacitor in the TTL Clock should be changed to 0.002 mfd, which speeds the sequence, so that filtering is effectively lower. Here we really have in mind that we want to go over to an external clock input anyway (a VCO) so our major concern is setting the filtering at a 16 kHz low-pass cutoff.

The circuit in Fig. 5 is a good white noise source, and can be used to drive the processing units of Fig. 2 and Fig. 3, replacing the noise source of Fig. 2. We are however using it mainly as an example, because if you are going to go to the trouble of building this more complex circuit, you will certainly want to add on some of the extra features we are about to describe.

The first feature we will consider is the external clock input. There are numerous circuits for interfacing an external clock to TTL. Here we choose the simple 555 timer shown in Fig. 6. The timer in this case is configured as a simple Schmitt

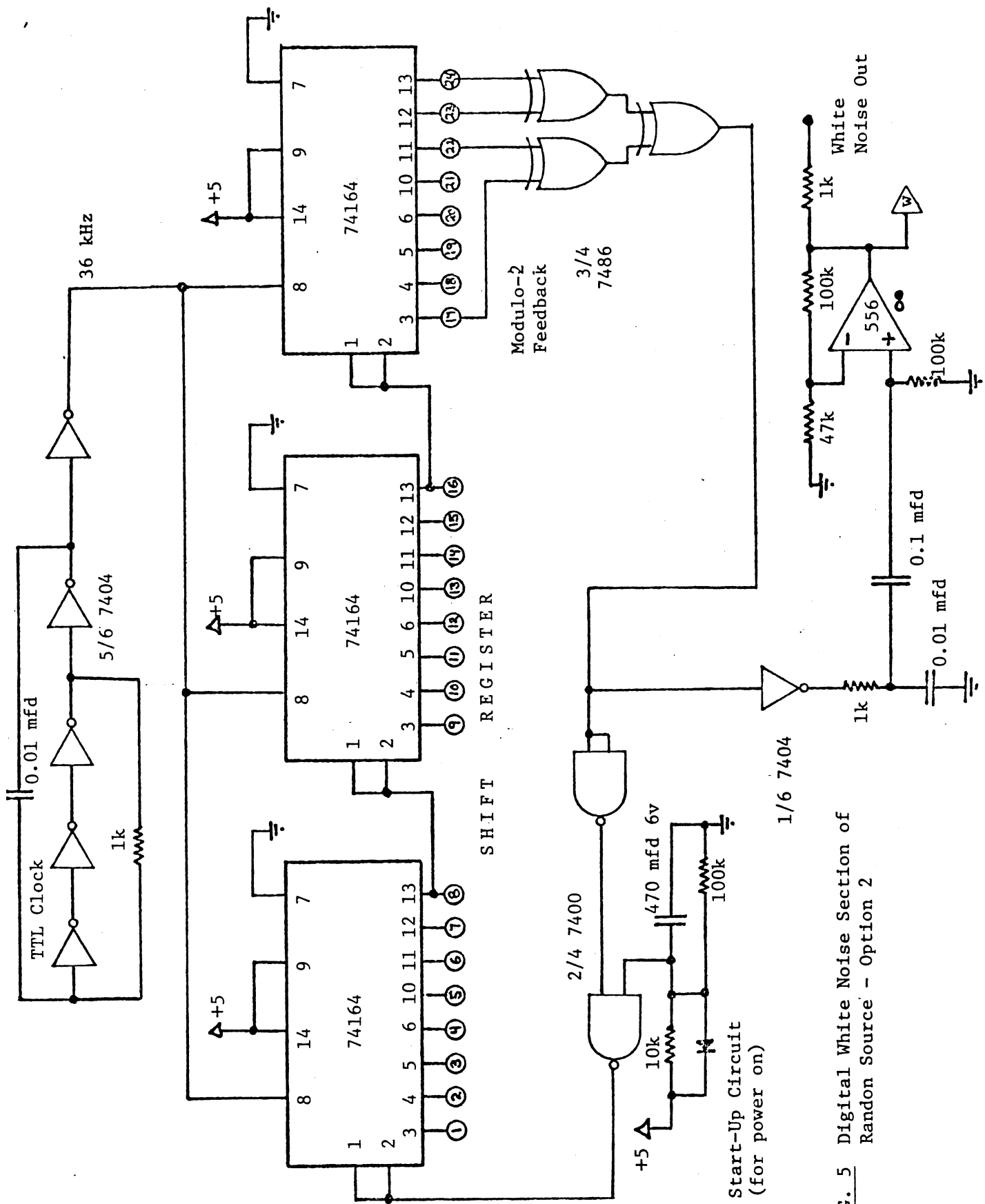


FIG. 5 Digital White Noise Section of  
Random Source - Option 2

trigger (See Application Note #31). This circuit requires that the input voltage pass through the region from 1.66 volts to 3.33 volts, which is the case for all our standard synthesizer waveforms. The Schmitt trigger action of the 555 means that we can use any waveform that is convenient from the external source - it need not be rectangular.

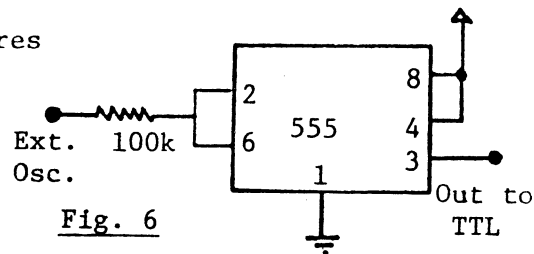


Fig. 6

The use of the external input with a quality musical VCO will become important when we consider the use of PR sequences as timbral elements. For the moment however, we note that with a wide range VCO attached, we can slow the sequence down into the sub-audio range, and the output can be used as a sequencer. Of course, if we want we can do the same general thing with a sample-and-hold on the full audio bandwidth output, but several interesting things happen when we try to use the lower clocking rates. First note that if we still wanted to use the sample-and-hold, we would have to use a much lower cutoff frequency on the low-pass filter, which is not really practical. Instead, it is practical to use the discrete time properties of the shift register itself rather than some external sample-and-hold. Thus we will consider groups of stages of the shift register to be digital words, and will convert these words to voltages using some sort of D/A converter. There are several ways to do this.

First, suppose we attach a standard 8-bit D/A converter to the first 8 stages of the shift register. We are interested in the different voltage levels that come out with each clock cycle. In particular, we inquire about the statistical distribution of levels and the expected changes in voltage between consecutive clock cycles. To do this properly, we have to know quite a bit about PR Sequences, but it will suit our purposes here to just know that on the average we can expect all possible combinations of 8 bits to appear in our 8-bit window with approximately equal probability. This means that all possible 8-bit numbers are equally likely, and the amplitude distribution is uniform. This is indicated in Fig. 7a. Thus, a binary weighted (standard) D/A attached to the shift register results in a uniform distribution. Next, we ask what would happen if we used equal weighting of all eight bits? The result is about the same as we would get if we shook up a box of eight pennies and counted all the heads equally. We expect that very often we get 4 or 5 heads, but cases where we get 8 heads, or only 1 head, are rare. The distribution is binomial, not uniform, and is a good approximation to the Gaussian distribution. Thus uniform weighting results in a Gaussian-like distribution (Fig. 7b). Finally we want to look at the variation of the output during consecutive clock cycles. One clock pulse shifts the bits of the shift register one place to the right. Thus in the case of uniform weighting, only the first and last bits will have any effect on the sum during any one clock pulse, and the maximum change is one bit. Thus, large changes are inhibited, and this means low-pass filtering of the sequence. If you are familiar with discrete time filters, you see that we have formed a low-pass transversal filter. Thus, the D/A connection implies some bandwidth reduction that must be allowed for, but this is of little importance for control sequences. Note also that with the binary weighted D/A there is also some low-pass filtering, although the exact function is different. Note further that if the taps for the D/A are not consecutive shift register positions, the degree of low-pass filtering is reduced, and larger changes are allowed (the maximum changes is allowed if there is at least one space between all taps). Note finally that if you have not understood a thing about this paragraph, you can find more information in EN#62, or you can read on for another page or so, and then come back, at which time your understanding may be increased by the examples we will consider.

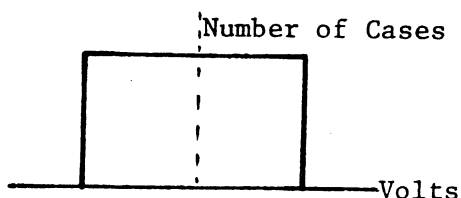


Fig. 7a Uniform Distribution

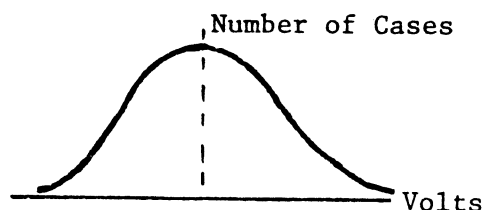


Fig. 7b Gaussian Distribution

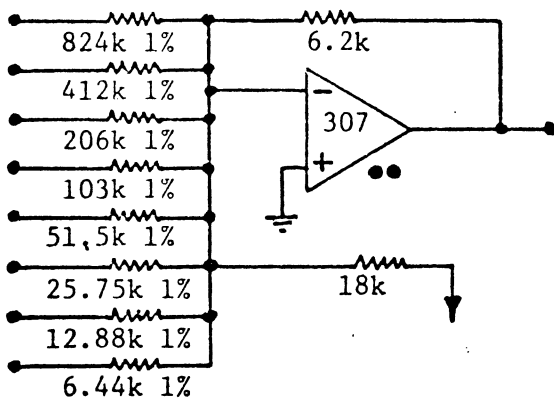


Fig. 8a Standard Binary Weighting  
Uniform Distribution

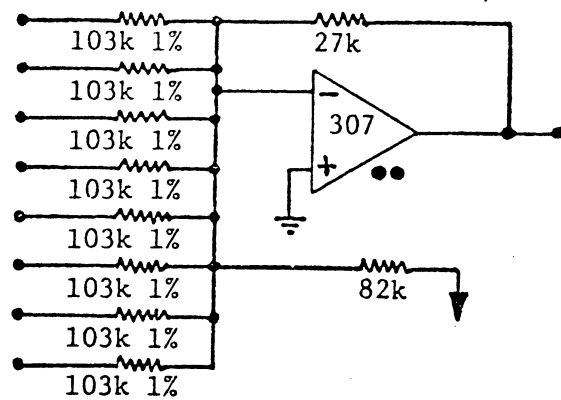


Fig. 8b Uniform Weighting  
Gaussian Distribution

Two circuits for D/A converters for the digital noise source are shown in Fig. 8a and Fig. 8b. These circuits should be interfaced with the 74164 using some sort of CMOS chips (inverters are fine - the inversion makes little difference with noise sources). The CMOS gives better defined digital levels. I used 74C00 chips for my interface, and the 103k resistors used just happened to be some I had on hand. You can scale the whole network for whatever you have on hand.

We have to consider here just what we expect to do with the random source to see what is required of the D/A's. In general, of course, we can not make 8 bit D/A's with 1% resistors (since 8 bits gives 256 levels). However, if the output of the source is to be used as audible noise, it is difficult to argue that errors caused by the 1% resistors will matter, since the effect of the errors is mainly additive white noise. Similarly, we would not insist that the op-amps in Fig. 2 be of the low noise type. What happens if we use the random source as a sequencer, driving a VCO for example. Ideally, the levels are equally spaced, and if we want 12-tone equally tempered scales, this gives about 120 levels over the audible 10 octave range. This is slightly less than 7 bits, and probably something like 6 bits or fewer is more like what we need for most musical purposes. We might therefore suppose that for sequencer operation, 6 bits could be used (see later design example), and that either the 1% resistors are close enough, or we could select, trim, or buy resistors of greater precision if necessary. If desired, the D/A converter used with the ENS-76 digital keyboard can be used in place of the one in Fig. 8a. This circuit is shown in EN#68 (16).

Next we want to consider how the taps on the shift register should be arranged, and the differences between the different outputs of the two different D/A's. We will look at this with regard to the effect these things have on the nature of the sequences that result, and we are generally thinking of the sequence as being converted to a sequence of pitches by a VCO, since this is a standard application. We can arrive at four cases for combinations of the two D/A's and close or wide spacing of the shift register taps. These four cases are diagrammed in Fig. 9. You can see that there are considerable differences between the four cases. These differences can be described in terms of the amplitude distribution and the degree of low-pass filtering, but we find it useful to combine these into one term which we will call "Expectation" which is a somewhat subjective feeling about what will come next in the sequence based on what has come before. It is related to our ability to find trends even in the absence of clear patterns. Narrow distributions like the Gaussian add Expectation because we do not expect extreme values. Low-pass filtering adds Expectation because we do not expect large changes. It is of some interest that the standard method of sampling white noise results in a moderate case of expectation, and is thus not as subjectively random as some users assume.

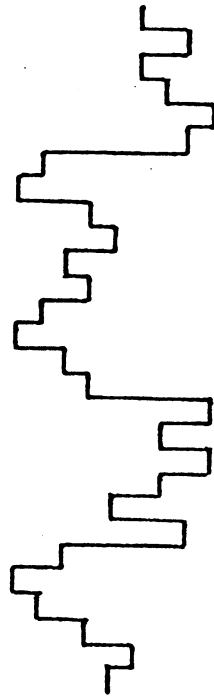
FIG. 9

BINARY WEIGHTING (Fig. 8a) UNIFORM DIST.

UNIFORM WEIGHTING (Fig. 8b) GAUSSIAN DIST.

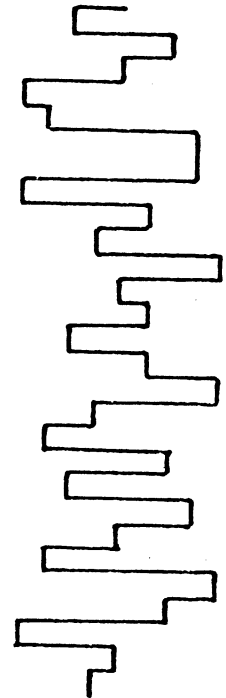
CONSECUTIVE  
TAPS ON THE  
SHIFT  
REGISTER

1. DISTRIBUTION: Uniform
2. LOW-PASS FILTERING: Some
3. EXPECTATION: Consecutive samples tend to cluster together, except when one of the more significant bits changes, at which time large jumps are possible.



MODERATE EXPECTATION

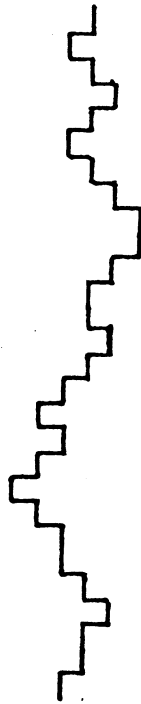
1. DISTRIBUTION: Uniform
2. LOW-PASS FILTERING: Little or None
3. EXPECTATION: None



LOWEST (NO) EXPECTATION

1. DISTRIBUTION: Gaussian
2. LOW-PASS FILTERING: Substantial
3. EXPECTATION: Large jumps are ruled out due to transversal low-pass filtering, and values far from zero are rare due to Gaussian distribution.

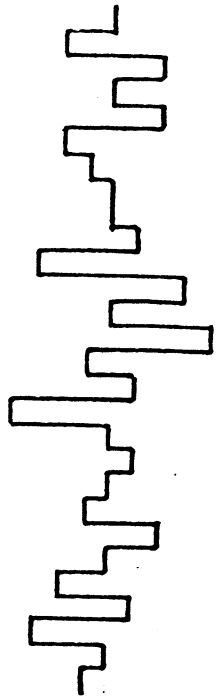
4. NOTES: Similar to Sample-and-Hold with white noise in and slew-limited S&H.



HIGHEST EXPECTATION

1. DISTRIBUTION: Gaussian
2. LOW-PASS FILTERING: Little or None
3. EXPECTATION: Large or small jumps are possible, but extreme values far from zero are rare due to Gaussian distribution.

4. NOTES: Very similar to process of using a standard Sample-and-Hold on a standard white noise source.



MODERATE EXPECTATION

WIDELY SPACED  
TAPS ON THE  
SHIFT  
REGISTER

There are no exact rules on how the taps on the line should be placed, even when we know which of the cases in Fig. 9 we are trying to get. As a start, the taps listed below can be tried. Probably you would want to have several different sets of taps available, so a 8-pole, 4-position switch, or the equivalent digital logic, could be used.

Close Spacing: 1, 2, 3, 4, 5, 6, 7, 8

Spaced Taps - 1: 1, 2, 3, 4, 6, 9, 13, 18

Spaced Taps - 2: 1, 6, 10, 13, 15, 16, 17, 18

Wide Spacing: 1, 3, 6, 10, 12, 15, 19, 21

Two different sets of "Spaced Taps" are given. These are combinations of consecutive taps and widely spaced taps. These give an in between case of Expectation. Two sets are given because it makes a difference with the binary weighted D/A.

## FURTHER OPTIONS

We want to discuss here some recycle devices that can be added on to the digital shift register for some new devices. First however, we should recall that when a PR sequence is of relatively short length, or even when it gets up to a length of a thousand or so, it still may be heard as a pitch when cycled rapidly. The exact reason for this is not completely understood, but probably the ear is not remembering the whole pattern and recognizing the repetition. Rather, the ear is probably recognizing the periodic cycling of some of the major "landmarks" in such a sequence. In any event, it is of interest that for short sequences we hear a sound that is not a great deal different than what we hear from a sawtooth or a pulse waveform. At the other extreme, we hear white noise for very long sequences. Clearly we are curious about what happens for medium length sequences. The area between random sounds and perfectly ordered sounds is little explored. There are many fascinating sounds to be heard for sequences from shift registers of between 10 and 20 stages. There are many more sounds available for non-maximal sequences. One can spend hours playing with a fully programmable sequence generator (see Mid-Month Letter 75A).

However, one thing can be concluded from a study of medium length sequences: We do not really get the in between sounds we were thinking we might get. Rather they sound like mixtures of random and ordered sounds - that is, noise and periodic clicks pops and dings. Interesting, possibly useful, but what else can we do?

Since short sequences result in interesting timbre generators (See R. Burhans, "Pseudo-Noise Timbre Generators," J. Aud Eng Soc. Vol. 20, No. 3, April 1972, pg. 3 and R. Burhans, "Harmonic Structure of PN Sequences," JAES Vol. 20) we can consider that short "captured" sub-sequences from longer sequences can also be used. This is easily implemented by adding an extra length of shift register to the output of the PR generator, and adding provisions so the contents of this added length can be loaded and then recirculated. The basic scheme should be evident from Fig. 10.

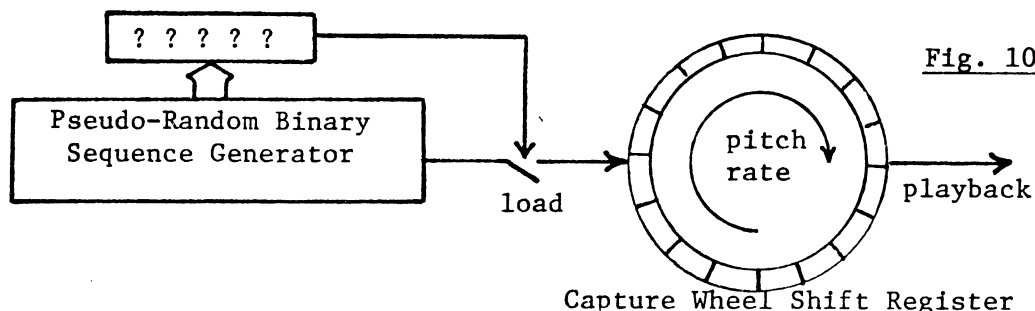
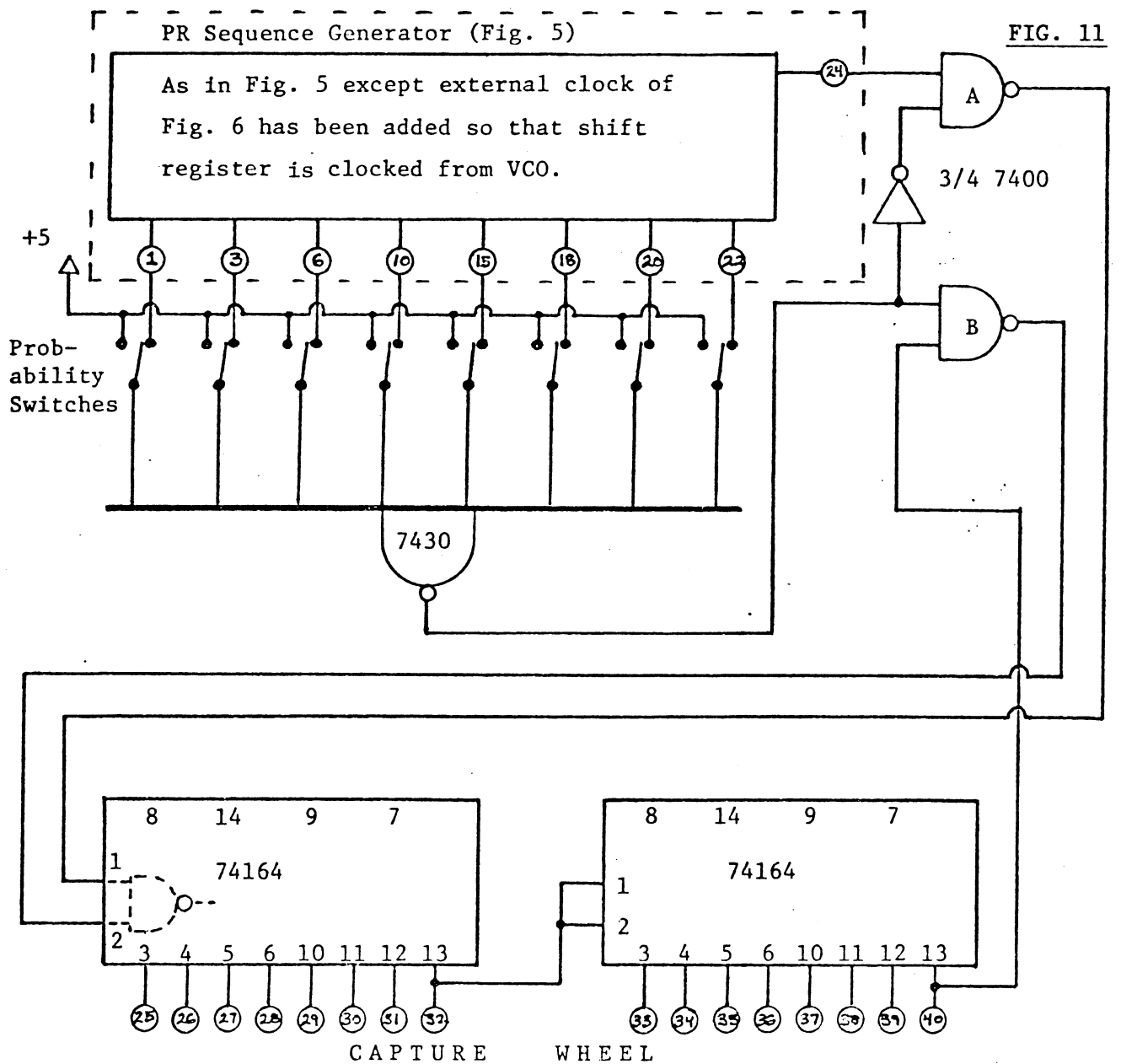




Fig. 10 shows a 16 stage shift register arranged as a "capture wheel" which can be loaded from the PR sequence generator. If we close the load switch, white noise is loaded on the wheel and read off at the other side. If we then open the load switch, the 16 states are fixed and one pitch cycle occurs on each rotation of the wheel. Thus we can easily obtain timbres somewhat at random, subject of course to the limitations of rectangular waveforms in general. We can also if we choose, periodically change one state of the shift register, or part of it by opening and closing the switch at regular intervals. However, by far the most interesting process is the one implied by the box with the question marks in it in Fig. 10. This permits the states on the wheel to be changed or not according to certain random (PR) conditions on the main PR sequence generator.

A TTL realization of Fig. 10 is shown in Fig. 11, where the PR Sequence generator is the one shown in Fig. 5. In normal operation, assume that the output of the 7430 8-In NAND gate is high so that NAND gate B is passing the sequence from stage 40 back through the input of the first of the 74164's in Fig. 11, thus recycling the sequence into stage 25 again. If the output of the 7430 actually goes low, then the recycle process is ended and new bits from the PR Sequence are input to stage 25. When will the output of the 7430 go low? When all eight inputs are high. This depends on the settings of the eight probability switches. Assume that two of the switches connect



the inputs to the +5 supply, while the other six switches are connected to the stages of the shift register. Now, any given stage of the PR Sequence generator may be high or low with equal probability. Thus, the probability that all six inputs are high is  $1/2^6$ . If this happens (on the average, once in  $2^6 = 64$  clock pulses), then the output of the 7430 gate goes low, and the contents of stage 24 are fed onto the capture wheel (to stage 25). The chance that this new state is different from the state currently on the capture wheel at that time is  $1/2$ , so the overall chance of a change of the particular bit on the capture wheel is  $(1/2^6) \cdot (1/2) = 1/2^7 = 1/128$ . Thus we could expect a change in one bit on the capture wheel for one clock cycle out of every 128, which for the 16 stage capture wheel is a change of one bit for every eight pitch cycles, on the average. Of course, this is only one example, and the probabilities can be changed by opening or closing more of the probability switches, and by adding more such switches if desired. Also, note that for the 16 stage capture wheel, the change of any one bit is a change of one part in 16, which is generally audible as a change of spectrum, but not always a major change. We can if we wish use fewer stages in the recycle wheel, or can add more shift registers. We suggest that as a starter, the 16 stage wheel works well, and a switch for changing from 8 to 16 stages might be useful.

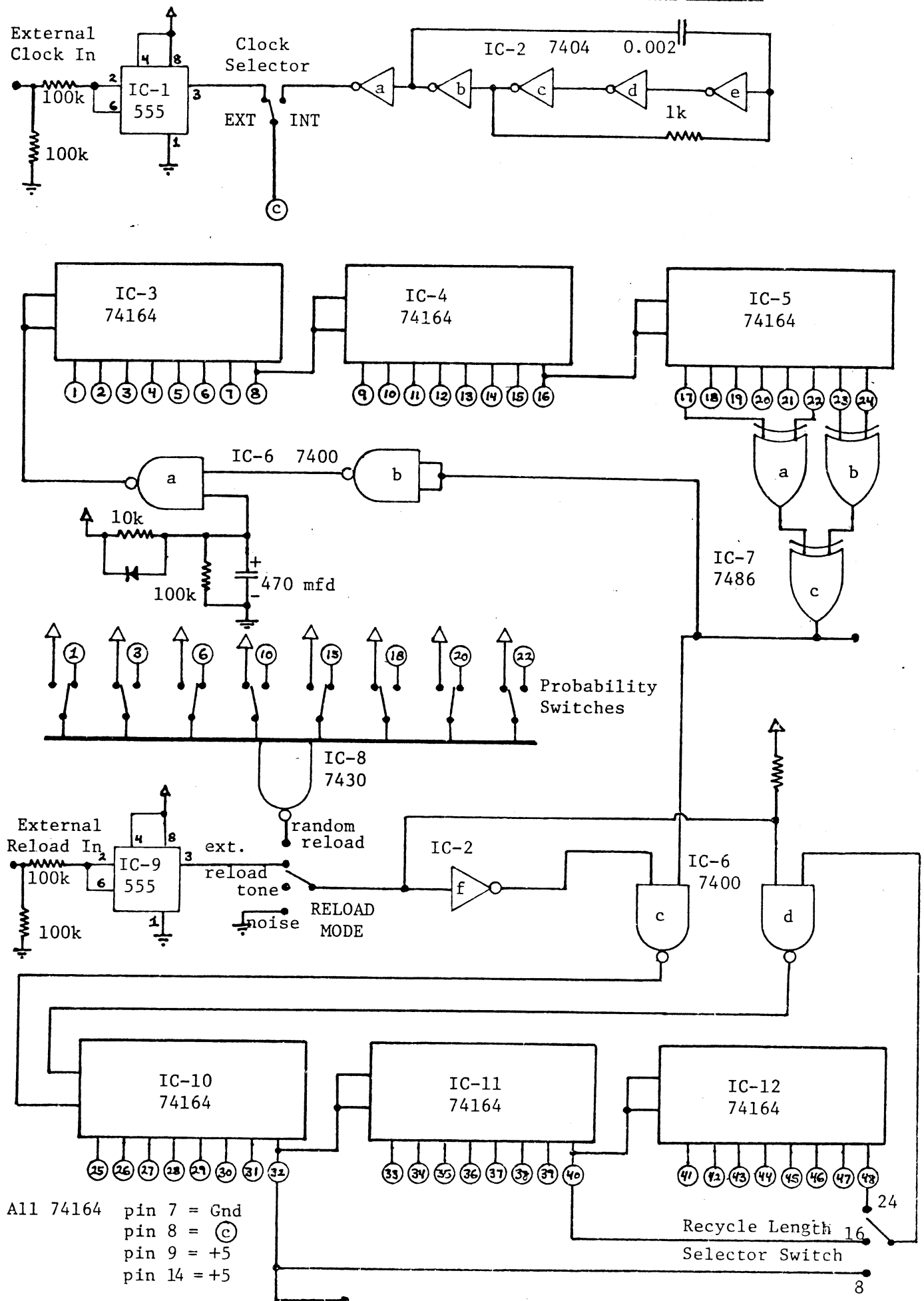
This process works quite well in general. It is felt that the sounds produced by this method do give sounds that are more like those in between random and perfectly periodic rather than mixtures of noise and periodic. What are the sounds like? Well of course it depends on the settings, since for certain probability settings the changes of timbre occur only once a second or so, while for others changes of timbre are quite rapid, and for the extreme of all switches connected to +5, we get white noise out. But in general, the sounds have a very broad spectrum, like a sharp pulse or a sawtooth waveform. Ralph Burhans (second reference mentioned above) showed that the PR sequences have a spectrum that includes all harmonics except those that are integer multiples of the sequence length, so the spectrum includes nearly all harmonics for practical purposes. Thus, the changes that occur when only one bit in 16 is changed may in general be expected to be subtle, although some are dramatic. Of course, the sequence can be further processed with filtering, and the VCO may be modulated as well for additional enrichment of the final spectrum.

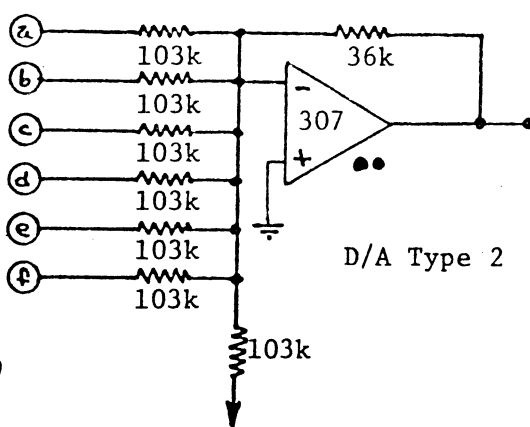
## DESIGN EXAMPLE

Since there are so many possibilities for a random source based on the principles given above, it is necessary to limit some of the features so that the resulting unit is not too complicated to be of use. The exact way this is limited is up to the individual designer, so we do not want to pin things down too much at this point. For this reason, instead of calling the following design Option 3, we prefer to call it just a design example.

The digital part of the design example is shown in Fig. 12 while the analog part is shown in Fig. 13. The circuit features either an internal clock (IC-2) or an interface for an external clock (IC-1). IC-3, IC-4, IC-5, and IC-7 form a 24 stage PR shift register with IC-6a,b forming the start up circuitry. IC-10, IC-11, and IC-12 form the recirculating or "capture wheel" part of the design. The length of the wheel is adjustable by the "Recycle Length Selector Switch" as 8, 16, or 24 stages. Note that the capture wheel is reloaded whenever a low state exists at the input of IC-2f, since this condition blocks the recycle through IC-6d, and feeds in PR bits from IC-7c through IC-6c. There are four positions on the "Reload Mode" switch. In the lowest position, the input of IC-2f is always low, and the PR sequence is passed onto the capture wheel at all times, and no periodic tone is produced. The second position (Tone) leaves the input of IC-2f always high, and whatever happens to be on the capture wheel at the time it goes high will be recirculated, thus forming a tone. Thus, the user can obtain somewhat random timbres by first loading noise (lower position) and then capturing and holding it (Tone position). There are also two automatic

FIG. 12 DIGITAL PART OF DESIGN EXAMPLE



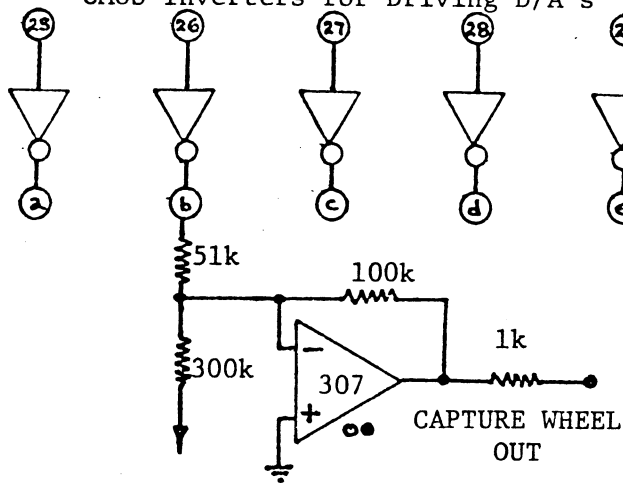


D/A Type 3

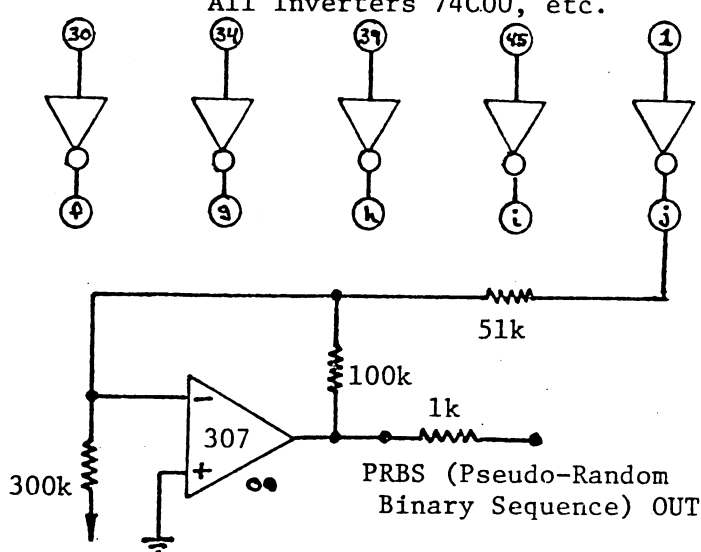
FIG. 1

The diagram shows a D/A Type 4 converter circuit. It features an operational amplifier (op-amp) labeled 307. The non-inverting input (+) is connected to ground. The inverting input (-) is connected to a ladder network of resistors. This network consists of seven 103k resistors in series, each connected to an input line labeled (a) through (g). The output of the op-amp is connected back to the inverting input through a 36k feedback resistor. The output line is labeled (h). Below the ladder network, there is a 103k resistor connected to ground, and the text 'PART OF LADDER' is written.

## CMOS Inverters for Driving D/A's



All Inverters 74C00, etc.



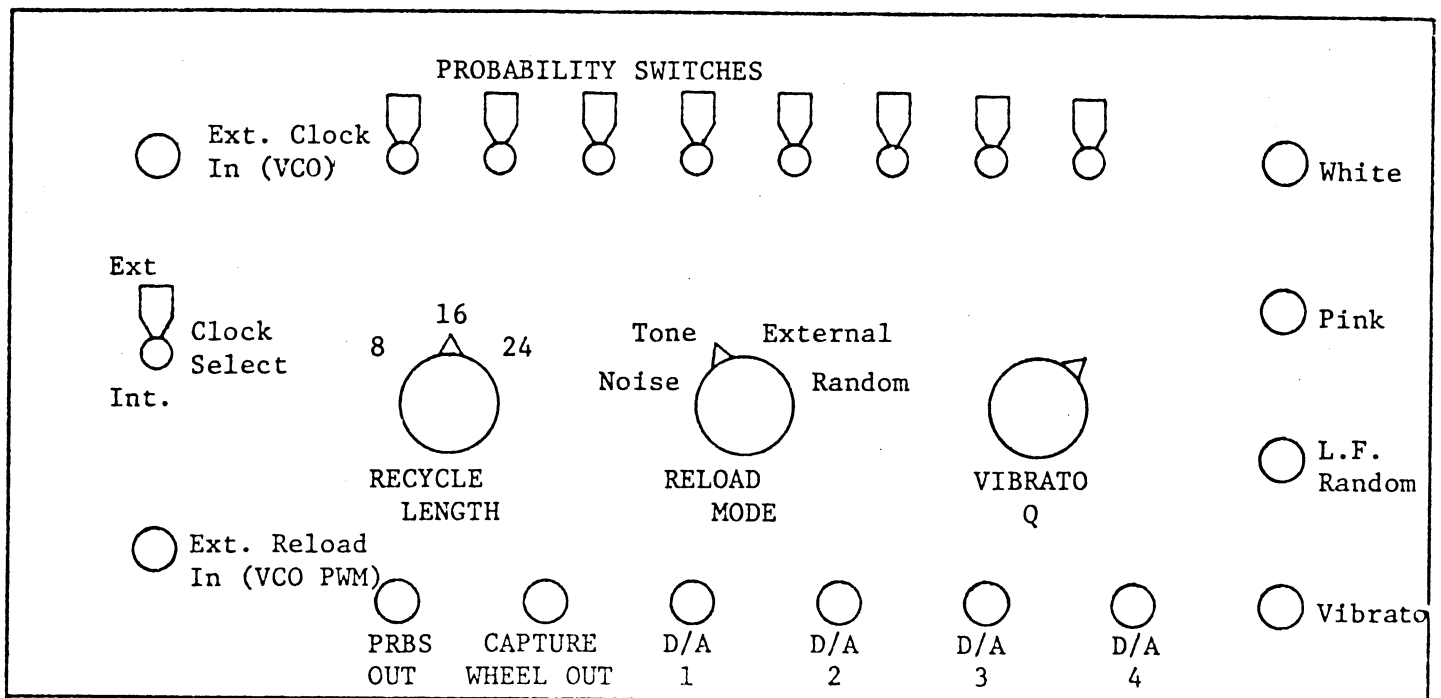
Before going any further, we want to try to head off any possible further confusion concerning the two shift registers. The first (IC-3, IC-4, and IC-5) is connected so as to always produce noise. It runs continually, and cycles after more than 16 million clock pulses. The reload function of the circuit in no way changes what is going on in this shift register. In particular, it does not change the length of the PR Sequence. We could obtain pitched sounds by shortening the PR Sequence and clocking it rapidly, but this is not what is done here. The second shift register (IC-10, IC-11, and IC-12)

while constructed in the same manner as the first, is really quite different in function. Its cycle length is never more than 24 clock pulses. If it is not in the recycle mode, it is simply a 24 stage extension for readout purposes only as far as the first shift register is concerned. It does not lengthen the PR Sequence because it follows the last Exclusive-OR gate, and there is no feedback from it to the first shift register. Once again, consulting Fig. 10 may be of help here.

The analog portion of the design example is shown in Fig. 13. Here we have made provisions to bring out different sequences, and have also added four simple D/A converters. Two of the D/A converters are uniformly weighted and two are binary weighted. We use two of each because we want to use both close and wide spacing of taps on each one, and it is really simpler to use separate D/A's rather than to work out a fancy switching arrangement. Also, we have four simultaneous outputs available. The D/A converters were cut back to 6 bits in this case to keep things simple. We have in mind that these will be used for random voltage sequences, but you can also take noise out of these if you wish. Note that the D/A's have been connected to the capture wheel shift register rather than to the PR Sequence shift register. This has been done so that captured sequences can also be read out through the D/A. By putting the reload mode switch in the noise position, we are effectively using the PR Sequence, only it is slightly delayed. Note that for wide spacing, some of the taps extend beyond IC-10 into IC-11 and IC-12. There is no real problem with this even when the recycle length switch is set to eight. In this case, IC-11 and IC-12 have the same contents as IC-10 but the spacing of the taps is such that no duplicated output is used for an input to a D/A. However, the fact that the sequence in IC-10 is duplicated in IC-11 and IC-12 has the effect of reducing the spacing of taps. This may be useful however since it has the effect of changing the spacing of the taps without actually changing them.

Things will perhaps be clearer if we look at a possible panel diagram for the random source design example. This is shown in Fig. 14. With clock select in the internal position, except for the Low Frequency Random and Vibrato outputs, which are similar to the outputs in Option 1. Let's now assume that an external musical quality VCO is input to the external clock input, and the external clock is selected. This would cause the bandwidth of the audio noise signals to rise and fall with VCO frequency. Now we listen to the Capture Wheel Out or any one of the D/A outputs, and switch the Reload Mode switch to tone. The output is a tone of high harmonic content. If we switch the Reload Mode

FIG. 14 POSSIBLE PANEL LAYOUT FOR DESIGN EXAMPLE



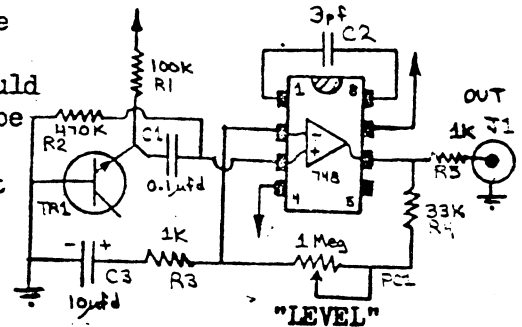
to noise and then back to tone again, we get a slightly different tone, and so on. Next we connect a second VCO to the External Reload input. We use a pulse-width modulated signal in here. We switch the Reload Mode switch to External, and adjust the second VCO to about  $\frac{1}{2}$  Hz and the pulse duty cycle to about  $\frac{1}{2}$ . We get out repeating bursts of noise 1 second long, and each of these noise bursts is followed by a tone of slightly different timbre. If the pulse width is changed so that it is nearly all high, we get mostly noise. If it is changed so that it is nearly all low, we get tones which change timbre every few seconds. This is because the external load signal is changing one bit on the capture wheel thus changing the timbre. If we switch the Reload Mode switch to Random, we get white noise out if all the probability switches are up (connected to +5). As we start setting the probability switches down one at a time, bits of tone start to appear and linger longer and longer. With all switches down, we have a tone of a constant pitch but one which is changing timber somewhat at random.

Next we can consider using the Random Source as a sequencer. For this, we can do all the same sort of things, but will be using the external clock at a very slow rate, probably 1 - 2 Hz, and will attach a VCO to one of the D/A outputs. With the Reload Mode switch in Noise, we get an apparently random tone sequence. Trying different D/A outputs will result in different degrees of Expectation. Now, here are a couple of points to consider if you want to use the Random Source as a sequencer. First, it might be well to replace the Recycle Length three position switch with a mulit-position switch with more terminals so that sequences of lengths other than 8, 16, and 24 can be selected. Secondly, you may want to add another position to the Reload Mode switch. This position would be fed by a counter working off the system clock. This counter could reload a bit every 13th clock pulse for example with a recycle length of 10, resulting in a sequential evolution at regular intervals. This would supplement the Random position where the sequence would evolve at random times. In either case, the exact notes of the evolving sequence are apparently random.

This is about all we want to say about this sort of device at this time. They are very simple to build, are not overly difficult to learn to use, give some very useful new capabilities, and work well with modular systems. Note in particular that the device is useful as both a sound generator and a sequencer, adding to its utility.

\* \* \* \* \*

ENS-73, NOISE SOURCE, OPTION 1: This is a very simple white noise source, useful for filtering with the VCF or to create random sequences with the S&H. IC-1 should be a 748 op-amp for best results, although a 741 can be used temporarily. TR-1 is really just an old noisy transistor. Try your old junkers, and use the noisiest one. A reverse biased zener diode can also be used. Upper noise peaks reach about 5 volts at the output. This circuit is from Dave Rossum for the ENS-73.



## NS-4

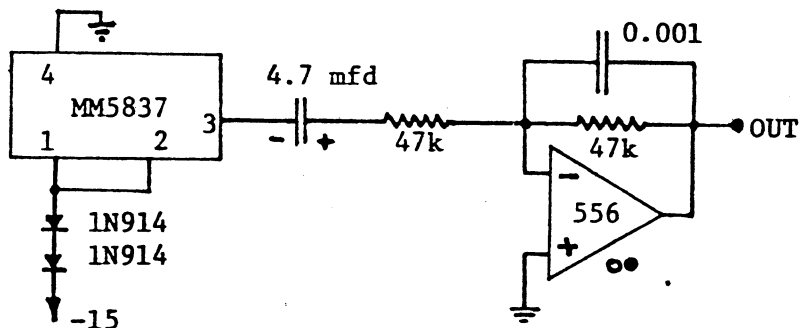


Fig. 11

## WHITE NOISE GENERATOR

The MM5837 is a 17 stage PRBS generator\*. It uses taps at stages 14 and 17, and has an internal clock of approximately 130 kHz. The 17 stage sequence is 131,071 bits long. Thus, the output at pin 3 is a PRBS that repeats about once per second. The amplitude of the PRBS is approximately 10 volts peak-to-peak. This is filtered by the single pole filter based around the 556 op-amp. We did not make any exact measurements on the output of the above source. Others may want to experiment with it more. We selected the 0.001 mF capacitor since it gave what appeared to be a Gaussian-like amplitude distribution, and still had a nice crisp sound to it. Others may wish to raise the low-pass cutoff frequency.

As we see it, this device has one outstanding advantage over the back-biased junction type of white noise source: it is a device independent design. With the back-biased junction, it was generally necessary to use a lot of amplification, and the exact amount depended on the individual transistor or diode used to generate the noise. With the above device, you know exactly what you are getting ahead of time. The major disadvantage is the periodicity of one second. As we mentioned earlier, this cycling is just audible, although not a real problem. A more important limitation is that when the noise source is sampled, the sampling frequency can not be set to a close ratio to the one second cycle time or else obviously periodic or near periodic sampling effects are found to occur. For example, a 1 Hz sampling rate will give a constant voltage at the output of the sample-and-hold. One solution would be to have a much longer cycle time by having more stages on the shift register. On the other hand, if the white noise is to be used primarily as a timbral element (as in bandpass filtering), the PRBS generator chip used above is a good choice.

EN#611





1 PHEASANT LANE

August 15, 1979

ITHACA, NY 14850

SIMPLE NOISE SOURCES

(607)-273-8030

Noise sources find application in areas of testing and for sound synthesis where their wideband spectra can be used to simulate a variety of noise sounds such as the wind or explosions. Noise sources can produce "white" noise (constant energy per unit of frequency) or "pink" noise (constant energy in a given frequency ratio), or any number of other spectral characteristics, either are a matter of their inherent spectrum, or through filtering. Noise sources are also characterized by their amplitude distributions (perhaps a Gaussian or bell-shaped distribution with extreme values unlikely, or a uniform distribution with all amplitudes equally likely, or by any number of other distributions, including a binary or two-level distribution). While it is useful to know that noise sources can be well characterized, this is not our concern here. Instead, we are interested in simple devices that will be adequate for many testing purposes, and for most sound synthesis purposes.

The first simple generator we want to discuss is based on the back-biased transistor junction as a noise source. Generally this is thought to be a wideband white noise, but it is of relatively low amplitude, and in general the bandwidth will be determined mainly by the limitations of the op-amps being used to amplify it. Thus, most such sources will have a much smaller bandwidth than is actually possible. Also, there is a drawback in production work that different transistors will have different amplitude levels, so each one may have to be individually tuned. By using two amplifiers (or more) instead of a single stage, a fairly wide bandwidth can be maintained. It may pay to vary the current in the transistor (by varying  $R_N$ ) slightly for a higher noise level.

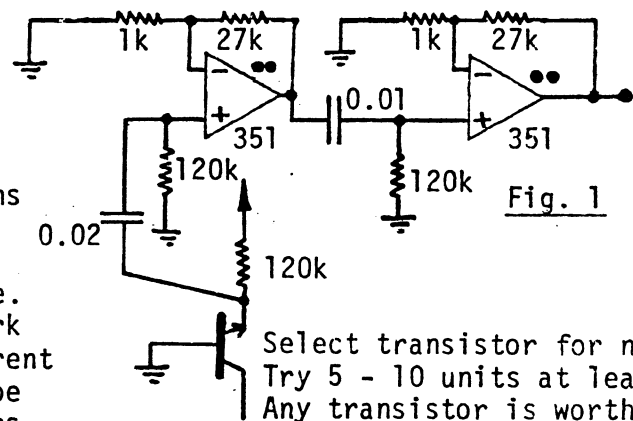
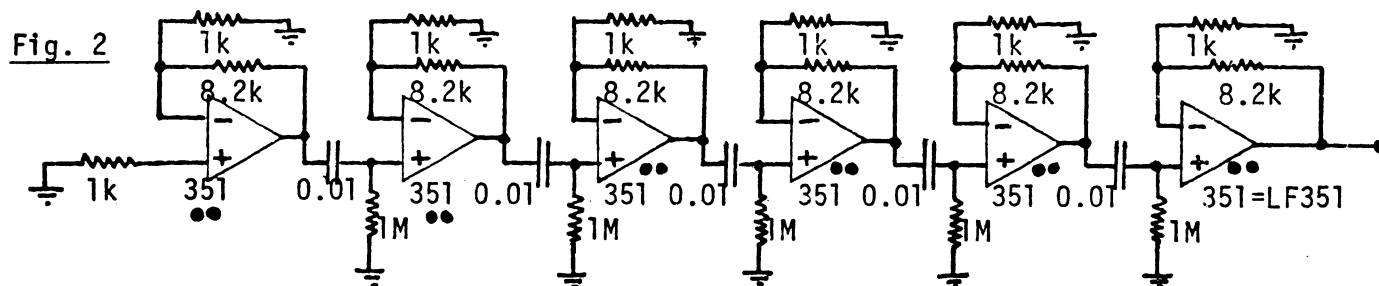


Fig. 1

Select transistor for noise.  
Try 5 - 10 units at least.  
Any transistor is worth a try

A simple noise source that has become popular lately is based on a method of just cascading op-amps, letting successive stages build up the level of noise present in the input stage. This is a useful method, and is fairly practical with op-amps becoming cheaper all the time (and you can use dual or quad units). Fig. 2 shows a cascade of six op-amps using this principle. In practice, the gain of all the op-amps is set so that all six are the same and so that the necessary amplitude level is reached. If



the resulting cascade does not give enough bandwidth, then more op-amps must be added and the gain decreased for each stage. Also, you can go the other way and reduce the number of op-amps if you need less amplitude or bandwidth. Note that you can do some bandwidth shaping by adjusting the gains of the different stages if you wish. Each of the op-amp stages provides a shaping to the spectrum according to the gain curve at the specified gain, and the internal 6db/octave roll-off due to the internal

compensation. That is, beyond the limits specified by the gain-bandwidth product, the spectrum is shaped at 6db/octave. This can be useful, or at least you can use it to estimate the bandwidth of your noise source.

A simple and well defined noise source is shown in Fig. 3 where a dedicated noise generator IC is the heart of the source. You should know however that the source is pseudo-random, not truly random. It is a digital two-level noise source, and repeats its sequence after a period of a few seconds. Being a digital circuit, we know exactly what the amplitude level is at the output, and it is relatively high, needing no real amplification, only buffering, and that is the purpose of the op-amp. The noise is limited at the output of the chip by the clock rate inside the chip. This rate is about 130 kHz and its inherent spectrum is  $(\sin x)/x$ , and cannot be expected to be very flat much above 50 kHz or so. The

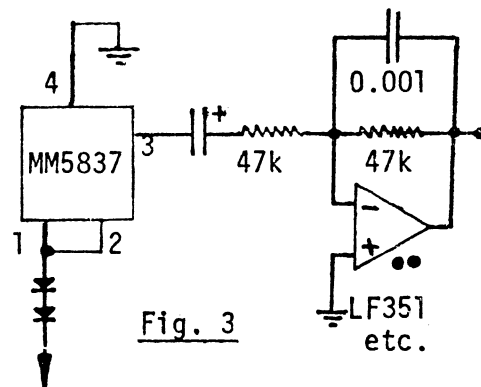


Fig. 3

op-amp is used mainly for buffering, and is configured as an inverter. This allows us to add a capacitor in parallel with the feedback resistor, and form a first-order low-pass filter. The cutoff frequency of this filter is  $1/2\pi RC$ , and is about 3.3 kHz as shown. The reason we set this so low is that we wanted not just filtering, but a spreading of the two level noise through the averaging effect of the low-pass filter. If too high a bandwidth is used, only two levels will be present in the output. The low-pass filter tends to average out several output values, just as flipping a box of coins will tend to give some heads and some tails, even though each coin has to be either a head or a tail. In the same process, extreme values tend to be rare, and we get more of a "Gaussian" distribution. Note that there is nothing particularly bad about a two level noise as far as the way it sounds goes. It still sounds like white noise, so if you need more bandwidth, and don't care about the amplitude distribution, you can greatly reduce the filtering, or remove it completely.

The IC of Fig. 3 is an example of the digital pseudo-random binary sequence generation technique, even though the details are not given. You can make your own PRBS generator using shift registers and one (or three in some cases) exclusive-OR gates. The basic idea is shown in Fig. 4 below:

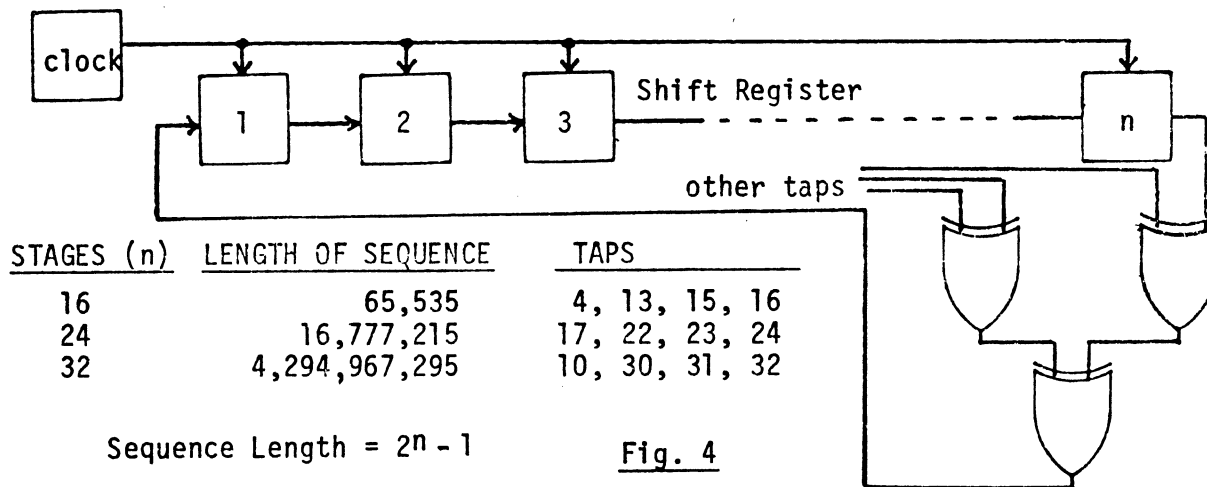
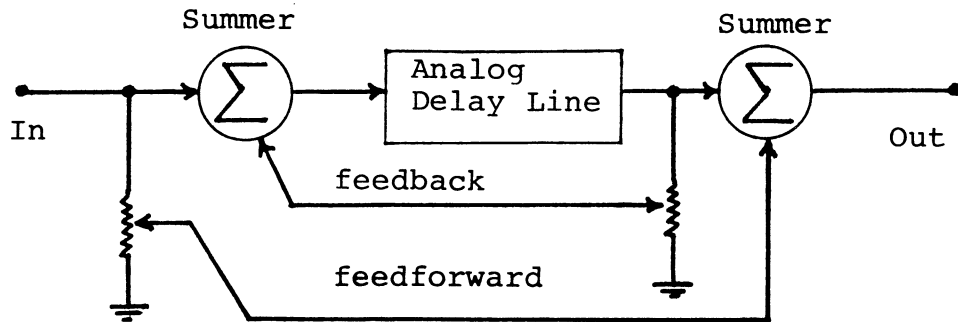


Fig. 4

PRBS generators are commonly formed from eight stage shift registers such as the 74164 (or 74C164 CMOS) and Exclusive-OR gates such as the 7486 (or 4070 CMOS). Since these come in units of 8 stages, the data in Fig. 4 is given for a total of 16, 24, and 32 stages. Note the increase in the length of the sequence as  $2^n - 1$  with n. Note that the 32 stage unit clocked at 1 MHz will not repeat for over an hour. Some care may have to be taken to assure that the generator does not start up in an all-zero state, as you can see that it will never get out of an all-zero state. If it starts in other than the all-zero state, the all-zero state will never occur [that's the -1 in  $2^n - 1$ , the one state that does not occur]. Of course, buffering, filtering, or whatever, can be done with this PRBS generator.

## ANALOG DELAY LINES

ANALOG DELAY LINES: Analog delay lines of the bucket-brigade type are part of a rapidly developing technology. We are interested in circuits that provide a variable delay time to an input analog signal. Typically such a delay is made part of a feedforward and feedback circuit as shown below, and is then called a "flanger" or a echo unit.



The first delay line option for this collection uses the Matsushita MN3001, a dual 512 stage delay line, and is the "Delay Module/Sub-Module" from the ENS-76 series. It features a voltage controlled clock, and two parallel delay lines with tracking VCF's on their outputs. As shown, it is intended as just a delay, but it can be put into the flanger circuit or used for a similar purpose. The second circuit designed by Jan Hall uses the 4096 stage MN3005 IC resulting in longer delay and overall better performance than some earlier chips. The third option is a delay evaluation board for the SAD-1024 from Reticon, and is somewhat similar to the first option of this collection. Note that the second option has the feedback option built in (recirculate).

| Preferred<br>Circuit<br>Option | Original<br>Source | Original<br>Designer | Estimated<br>Parts<br>Cost | Notes            |
|--------------------------------|--------------------|----------------------|----------------------------|------------------|
| ADL-1                          | EN#72              | Hutchins*            | \$30                       | ENS-76 series    |
| ADL-2                          | EN#87              | Hall                 | \$58                       | longest delay    |
| ADL-3                          | AN-35              | Hutchins             | \$33                       | Reticon SAD-1024 |

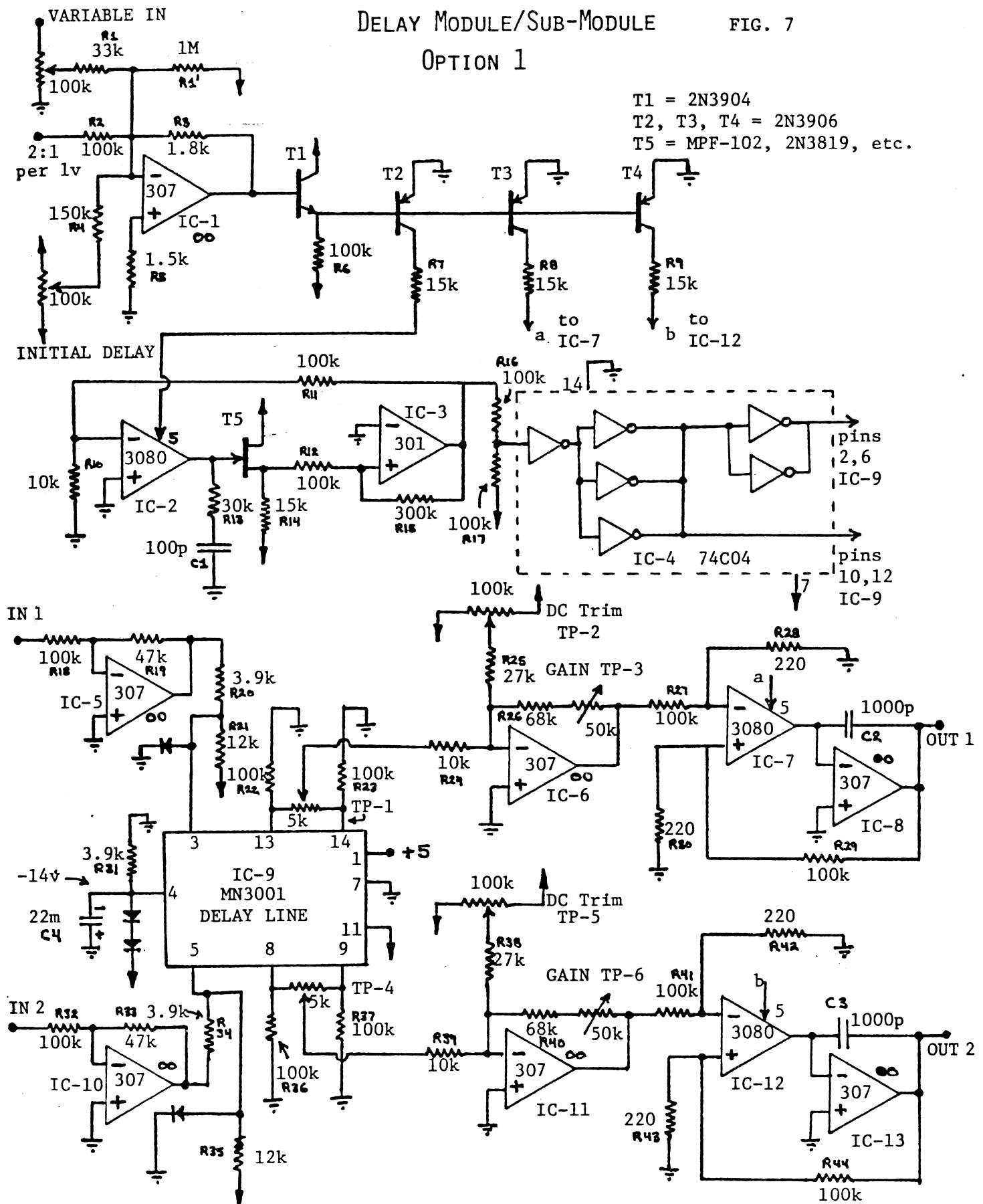
\*based in part on an earlier design of Jan Hall

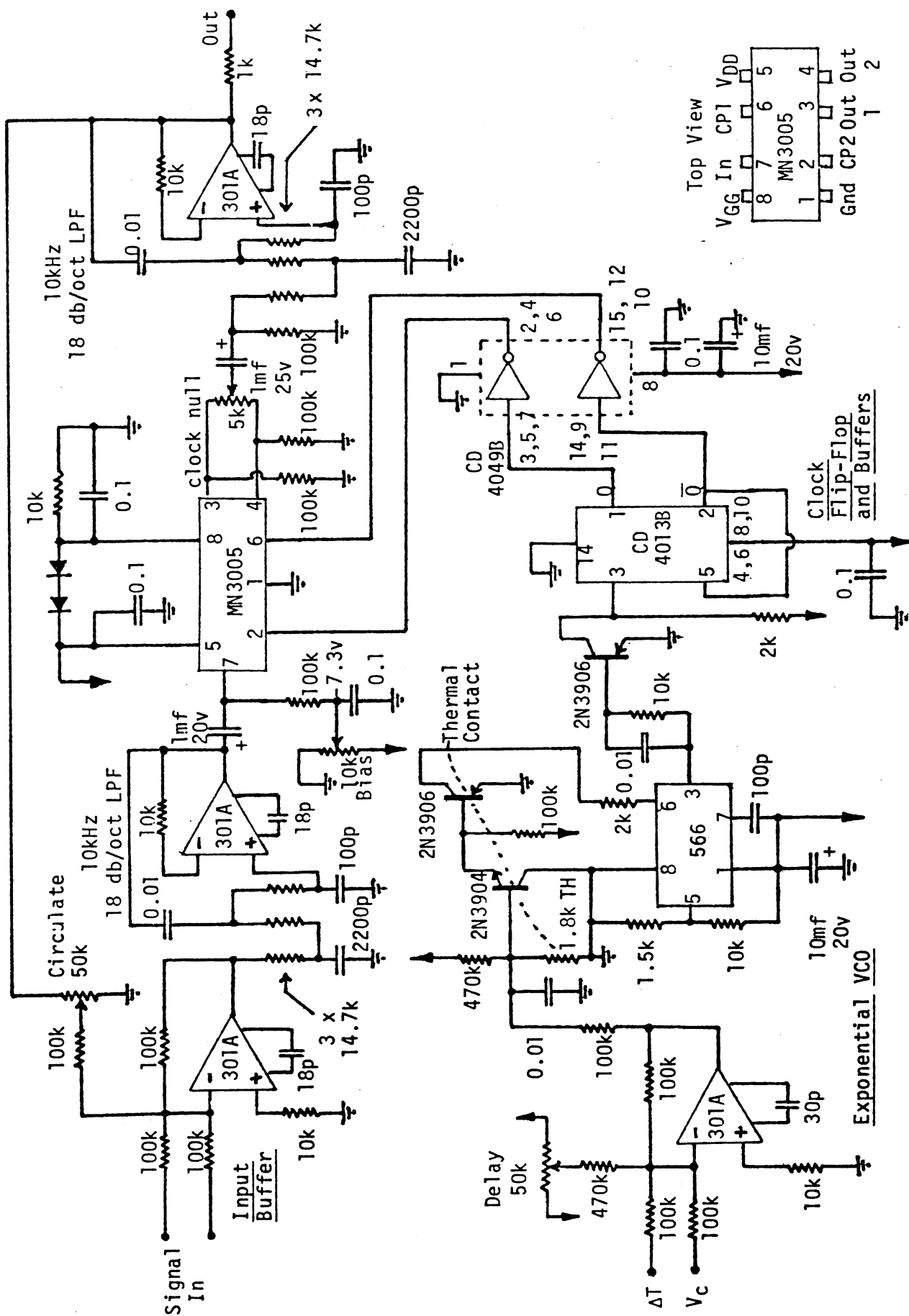


# DELAY MODULE/SUB-MODULE OPTION 1

FIG. 7

T1 = 2N3904  
T2, T3, T4 = 2N3906  
T5 = MPF-102, 2N3819, etc.





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APPLICATION NOTE NO. 35

April 18, 1977

DELAY LINE SETUP USING THE SAD-1024

In AN-34, we described a delay line setup using the Matsushita MN3001 bucket brigade chip. Here we will describe a very similar setup using the Reticon SAD-1024, which is a functionally equivalent device. A quick glance at the circuit diagram on page 2 of this note will show that much of the circuit has been copied directly from AN-34. There are a few changes that had to be made to change from the MN3001. Also, there are a number of things in the present SAD-1024 circuit that can be carried back to the MN3001 circuit if you wish. In any case, we suggest that you read or review AN-34 before going on to this one.

The SAD-1024 has somewhat simpler power supply requirements than the MN3001. It need only a total supply of 15 volts. If we wished, we could run it on ground and -15 as with the MN3001, and very few changes would be necessary in the support circuitry. However, since we require slightly less attenuation in the DC coupler if +15 and ground are used, we shall choose these levels, and use much of the circuitry of the standard Reticon evaluation board from there on. The first step is to change the power supply lines from the CMOS flip-flop clock so that they range from ground to +15. In the circuit of this note, we also change to a D-Type flip-flop as well, and parallel both flip-flops in the 4013 chip, and parallel the input lines to the SAD-1024 clock pins. We could if we desire use separate flip-flops for each section of the dual 512 stage chip, or use the 4027 flip-flop from AN-34. We see little reason to prefer one method to any other.

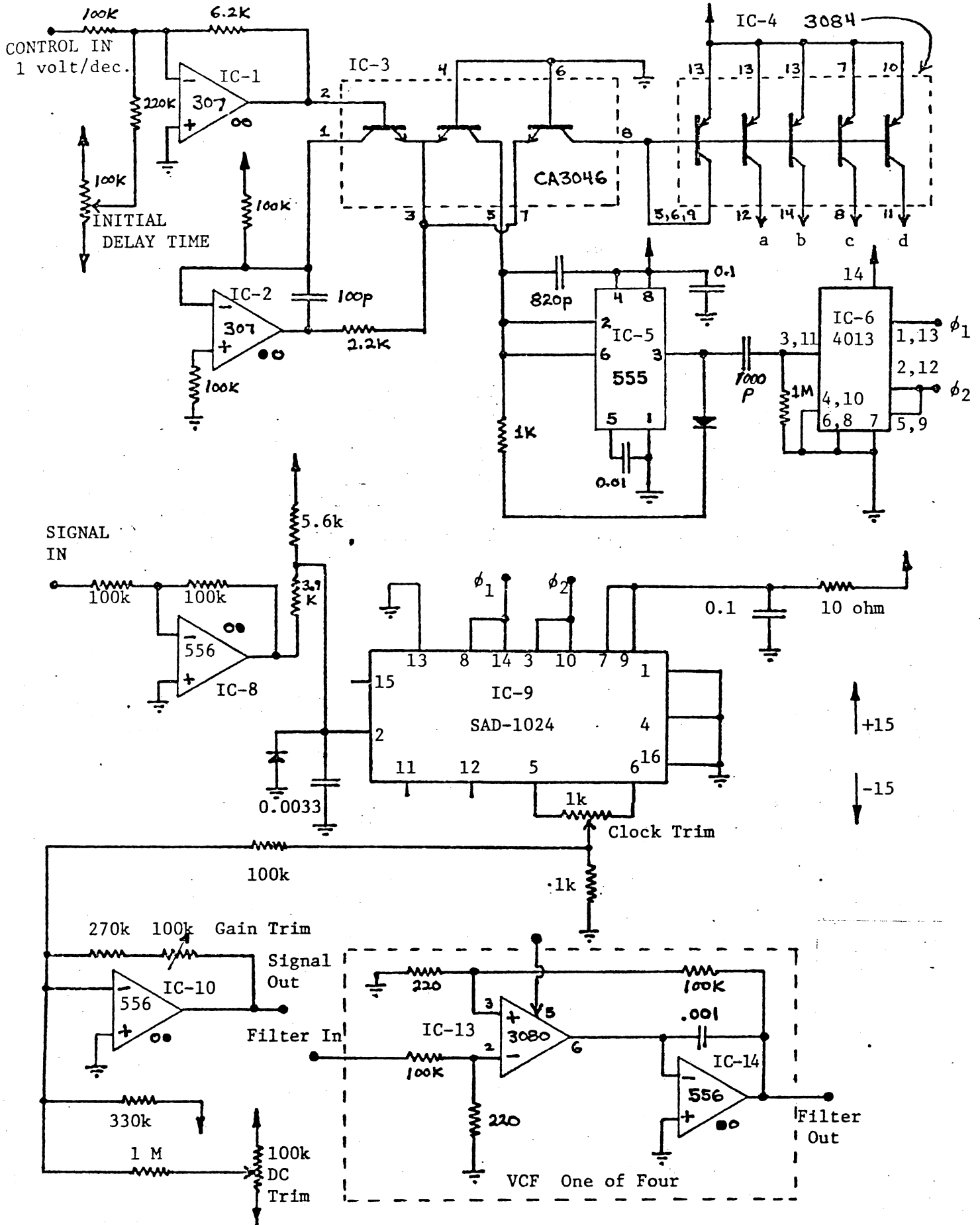
The SAD-1025 also needs a DC operating level of about +6 volts rather than the -3.7 volts of the MN3001, so this time the voltage divider from IC-8 goes through a 3.9k-5.6k divider to +15 supply. As before, the passive guard for a 20 kHz single-pole low-pass cutoff is used, 0.0033 mfd in this case.

We also use decoupling of the +15 power supply using the 10 ohm resistor and 0.1 mfd capacitor as shown, as in the Reticon evaluation circuit. The output circuit, like the MN3001 uses a clock trimmer. This time however, the trimmer is a 1k type, and instead of two resistors to ground, there is only one here (1k). The final task is to DC level shift the output of the SAD-1024 back to ground, and set the gain so that the gain is unity from signal input to signal output. The gain is easiest set using an AC voltmeter on the input and output. Once the gain is set correctly, the DC level can be set to zero. Note that in the MN3001 circuit we used gain trim in the input resistor of IC-10, while here we use gain trim in the feedback loop. This is just by way of example, as it should make little difference in either case.

We also show as before, one of four of the tracking low-pass filters (IC-13 and IC-14). Not shown are the other three filters formed around IC-15 to IC-20. The second line is set up the same way as the first of the 512 stage lines. An input stage, based around IC-11 (not shown), is identical to the circuitry around IC-8, and attaches to pin 15 of IC-9. The second output stage, based around IC-12 (not shown) is identical to the circuitry around IC-10, and attaches to pins 11 and 12 of IC-9.

The reader should be sure to note the difference between the passive guard filter (the 0.0033 mfd capacitor) and the tracking VCF's, which may serve as guard filters. The passive guard is permanently in place, and will help to remove any signals above 20 kHz, which otherwise would enter the discrete time system. This is used here to keep noise out, thus preventing high frequency noise from being aliased down to the audio range. The tracking VCF's are set according to the needs of the application.

# DELAY LINE EVALUATION SETUP





## ANIMATORS

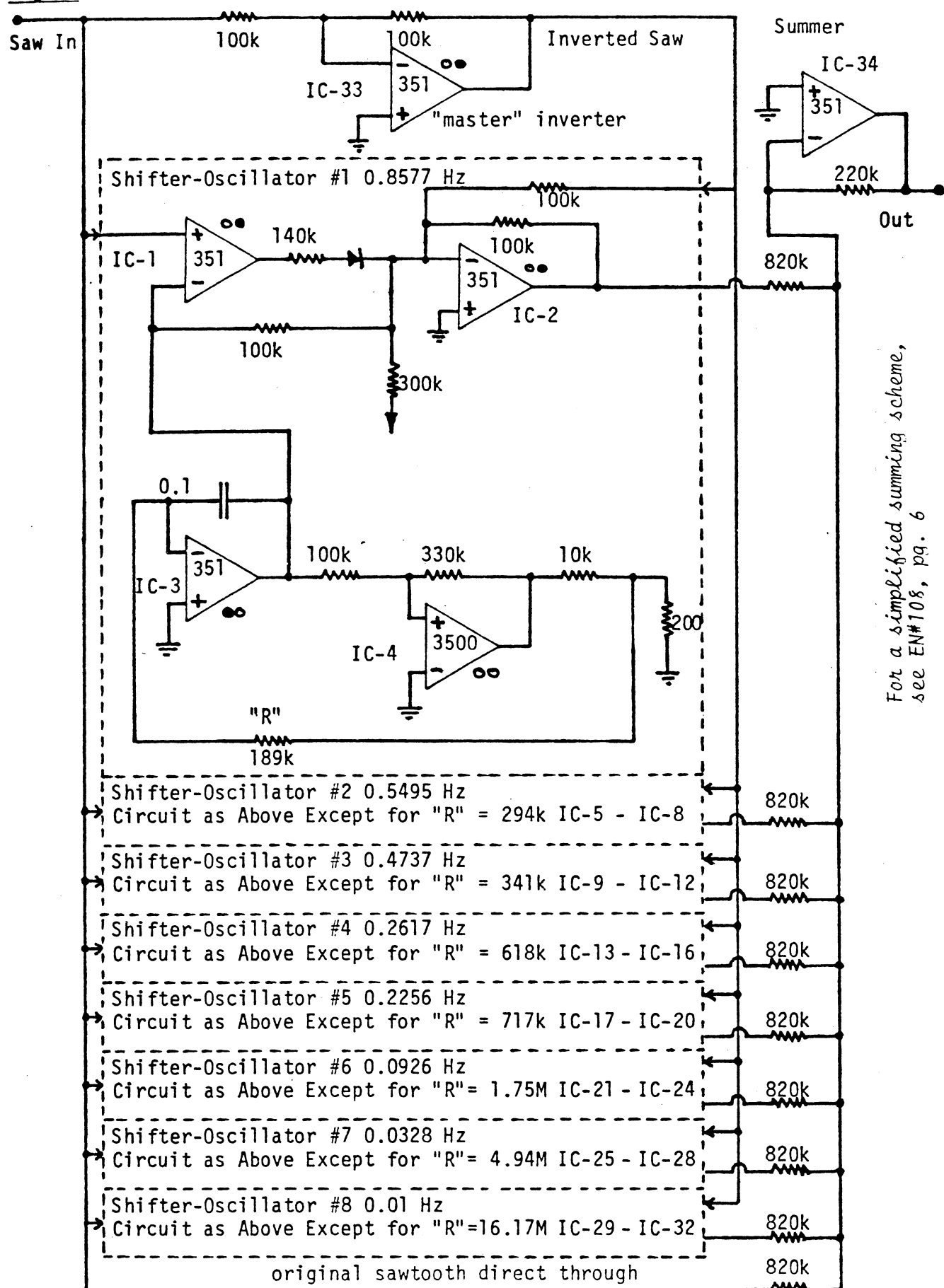
Animators: Animators are devices which enrich a waveform from an oscillator, and usually are intended for steady state portions of tones. This collection contains two animators. The first is a "Multi-Phase Waveform Animator" (MPWA) that first generates additional phases from a sawtooth (or other waveform) and then sums these all together, while at the same time the phase relationships gradually evolve. The second animator is a "Vocal Effects Waveform Animator" (VEWA) which creates a moving formant filter with the formant positions and motions following a vocal track model to some approximation. Both are successful animators, but applications are still experimental at this time. Additional animators appear in Electronotes and more are planned. The animators are:

| <u>OPTION</u> | <u>SOURCE</u> | <u>DESIGNER</u> | <u>APPROX COST</u> | <u>NOTES</u> |
|---------------|---------------|-----------------|--------------------|--------------|
| ANM-1         | EN#87         | Hutchins        | \$24               | MPWA         |
| ANM-2         | EN#102        | Hutchins        | \$28               | VEWA         |



Fig. 6 Full Circuit of Multi-Phase Waveform Animator

ANM-1



Notes: Op-Amp IC-4, etc., marked 3500 may be BB3500, 307, 741, etc.  
 All "R" resistors may be closest 5% value to value listed.  
 For different operating frequencies, choose "R" from Table 1

Fig. 3 VC LFO's and Summers

ANM-2

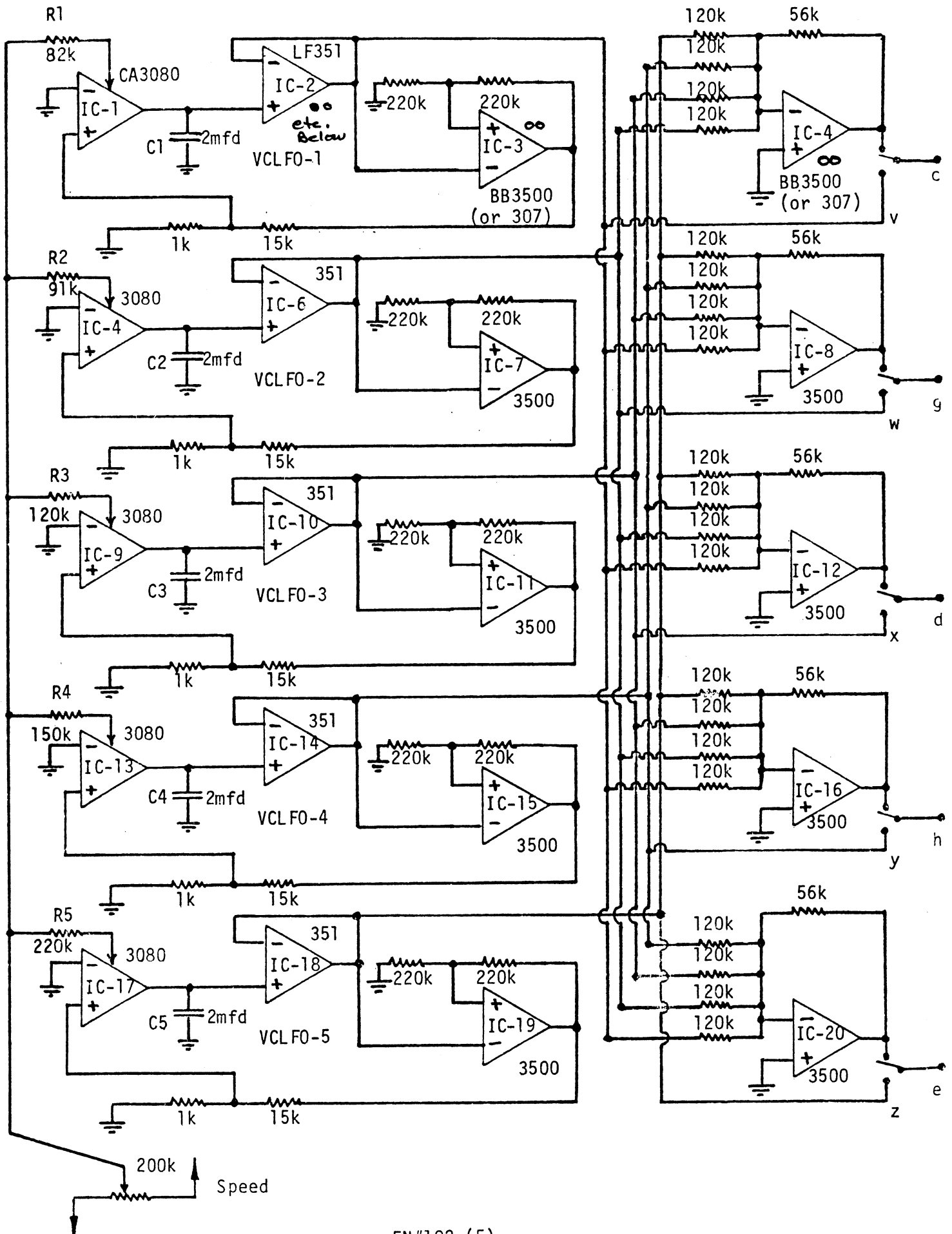




Fig. 5 Notch

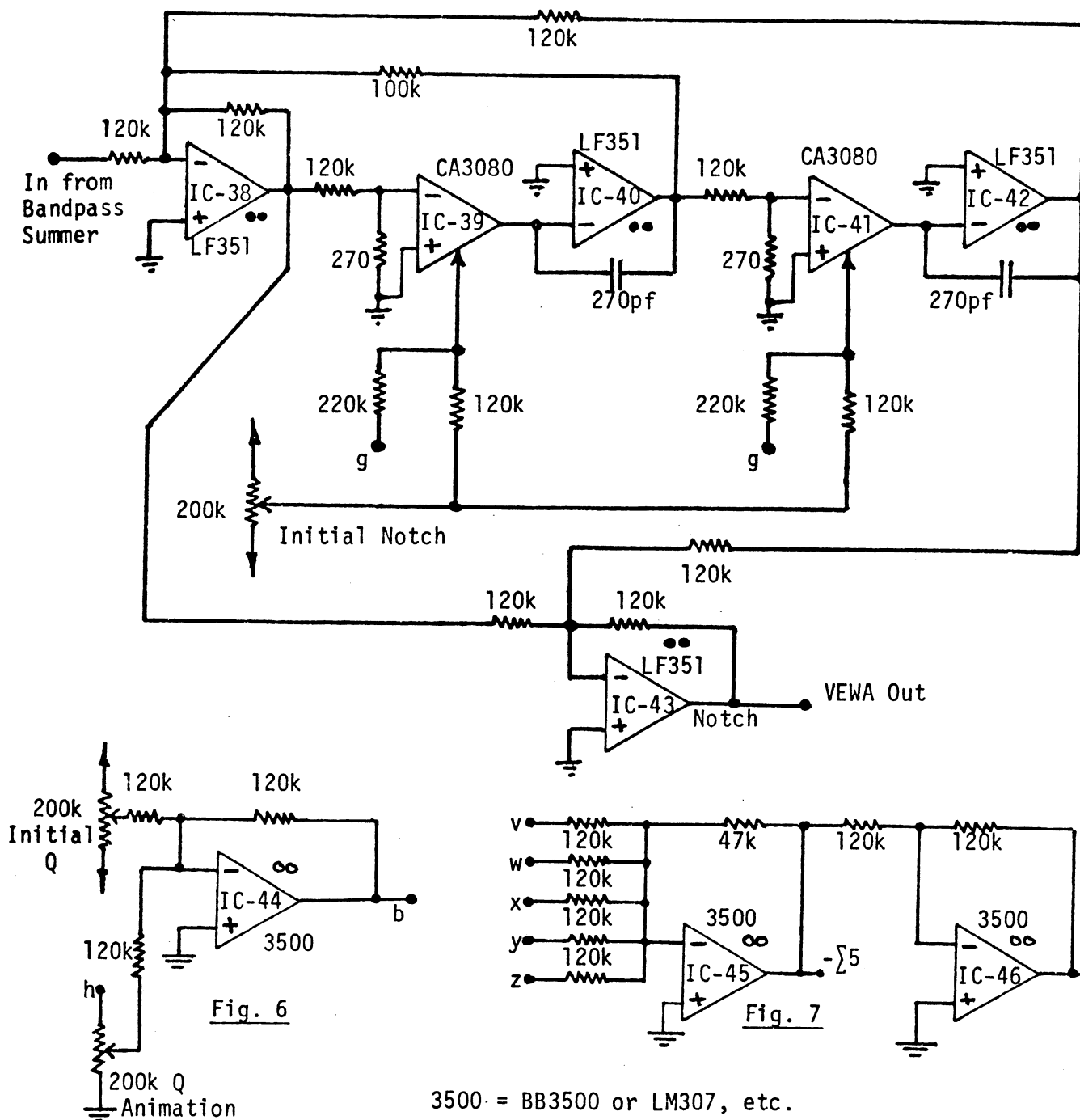


Fig. 5 shows a fourth state-variable VCF, this one without voltage-controlled Q, but with a notch feature. Here it is necessary to have a high-pass summer because we need to sum low-pass and high-pass to get the notch we desire. The result is a fairly conventional design. Note that the notch VCF follows the bandpass summer and is controlled by its own frequency level control (notch position) and by an animating voltage.

Fig. 6 and Fig. 7 are basically left-over circuits that did not fit well on the other diagrams. Fig. 6 is the Q-control summer. Note first the effect of the initial Q control. As the voltage on the wiper of this pot gets more positive, the voltage at the output of the summer gets more negative. Thus the control currents to the CA3080 Q-controls will get smaller, and this means less feedback, and therefore a higher Q. Thus, more positive voltages result in a higher Q. In addition to the initial Q, an animating voltage can be superimposed, and to a degree determined by

# PART FOUR

## APPENDICES

- A

 BUT WHAT SHOULD I BUILD RIGHT NOW?  
Reprinted from EN#91
- B

 AIDS ON OBTAINING PARTS
- C

 TROUBLESHOOTING  
App. Notes 131-134
- D

 INFORMATION ON SPECIAL PURPOSE  
ELECTRONIC MUSIC IC'S
- E

 USEFUL REFERENCES





BUT WHAT SHOULD I START BUILDING RIGHT NOW?

To answer this question, we would probably have to ask you a dozen more questions first as to what your resources and goals are. For this reason, we will be answering this after first setting up some examples or "scenarios" into which you may be able to place yourself, or at least to get some guidance.

▷ SCENARIO 1: The "Electronic Music Freak"

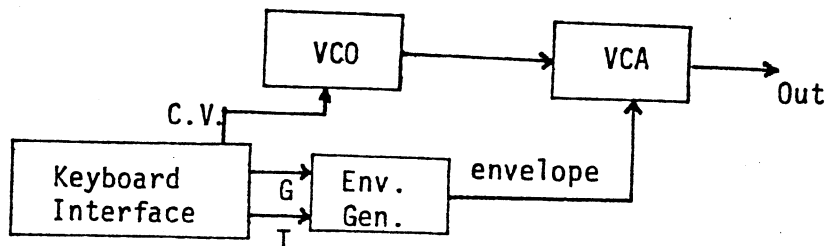
In this scenario we have your typical electronic music freak - someone who is for the moment into electronic music for its own sake, for the fun of it. He would like to have had his system running yesterday! We further assume the following goals and resources:

|                                     |                               |
|-------------------------------------|-------------------------------|
| Musical Goals:                      | Modest                        |
| Engineering Goals:                  | Possible Career               |
| Immediate Budget:                   | \$200                         |
| Expected additional funds (1 year): | \$200                         |
| Time Requirements:                  | None except for personal zeal |

For this person, we will choose modules that will get the system "on the air" as soon as possible. This will be similar to the "minimum operational" synthesizer we listed some time ago. The modules selected will be:

|                    |                  |
|--------------------|------------------|
| Keyboard Interface | Digital Design   |
| VCO                | ENS-76, Option 2 |
| VCA                | ENS-76           |
| Envelope Generator | ENS-76, Option 1 |

With the cost of the keyboard, these four devices will use up about \$150 of the immediate budget (including power supply), but you will be able to play the setup using the patch below:



The remainder of the immediate budget then goes into a VCF (ENS-76 Option 1 suggested), a second VCO of the same type, a second envelope generator of the same type, and probably a balanced modulator (ENS-76) and a sample-and-hold (from EN#61). At this point, the builder has a reasonable starting system, and can plan his expansion for the second \$200 of his budget over the next year. With the system so far, he will have a much better idea what he wants to add next.

▷ SCENARIO 2: The Professional Studio

Here we have a setup going into a professional studio. We assume that there is recording equipment already there, so any funds mentioned will be for the synthesis equipment. Further, there is no synthesis equipment in the studio. The following goals and resources on the part of the builder are assumed:

|                    |                           |
|--------------------|---------------------------|
| Musical Goals:     | Professional Recording    |
| Engineering Goals: | Construction, Maintenance |

|                                     |                             |
|-------------------------------------|-----------------------------|
| Immediate Budget:                   | \$400                       |
| Expected additional funds (1 year): | \$600                       |
| Time Requirements:                  | Operating in 1 year maximum |

In this case, there is a little more time and money involved, so we will be using the better circuits and probably taking more time to make things look neat, as this may be important for a professional studio. Probably the comments here will also apply to the individual builder who prefers to have a little better equipment and is willing to invest more time and money into it. We will choose the following modules:

|                                   |                              |
|-----------------------------------|------------------------------|
| Keyboard Interface                | Digital Design               |
| VCO's (three units)               | ENS-76 Option 1              |
| VCA's (two dual units)            | SSM-2020 design              |
| VCF's (two different units)       | ENS-76 Option 1              |
|                                   | SSM-2040 Four-pole design    |
| Envelope Generators (three units) | ENS-76 Option 2 or           |
|                                   | SSM-2050 design or mix       |
|                                   | of these two designs         |
| Balanced Modulators (two units)   | AD533 design                 |
| Sample-and-Hold (two units)       | EN#61 design                 |
| Noise Source                      | ENS-76 (analog with access.) |

This will require approximately \$300 of the available first \$400. The remainder will go into additional modules such as timbre modulators and slew limiters, and certainly for a professional studio, a lot will also go for fancy panels, fancy knobs, nice patch cords, and so on, and this may well run over into the rest of the first year funds.

The second phase of the construction, as in the first scenario, will depend in large part on what is learned from working with the modules constructed so far. Some logical devices to add during the second stage would be additional VCO's, VCF's, and more envelope generators. Also, new capabilities will be added by building the frequency shifter, one or more MPWA (timbre enhancement devices), the variable slope VCF, one through-zero VCO, and an analog delay line. All this equipment, well built and in good repair, would certainly make a nice studio.

### ► SCENARIO 3: The Professional Bandsman

In this scenario, we have the professional touring musician who for reasons of his own prefers to build his own synthesizer rather than buy one of the many small synthesizers aimed at this particular market. The following goals and resources are assumed:

|                                    |                           |
|------------------------------------|---------------------------|
| Musical Goals:                     | Performance professional  |
| Engineering Goals:                 | Maintenance and curiosity |
| Immediate Budget:                  | \$500                     |
| Expected additional funds (1 year) | \$200                     |
| Time Requirements                  | 4 months to road          |

In a first approach to this problem, you might just look at the commercial unit that comes closest to your needs and copy its features, using our most appropriate circuitry. Here we will be doing something a little different. The first thing

to consider is that this type of performance generally requires a high level of capability in the "imitative synthesis" area - we have to be able to imitate the sound of traditional instruments, not exactly, but in terms of their essential features, which means the manipulation of upper harmonics during the progress of a single tone. To do this we have available our VCF's, but here we also look to dynamic depth FM at the same time. The modules selected are:

#### Keyboard Interface

VCO's (three total)

VCF

VCA

Envelope Generators (three total)

#### Digital Design

ENS-76 (option 3) [one]

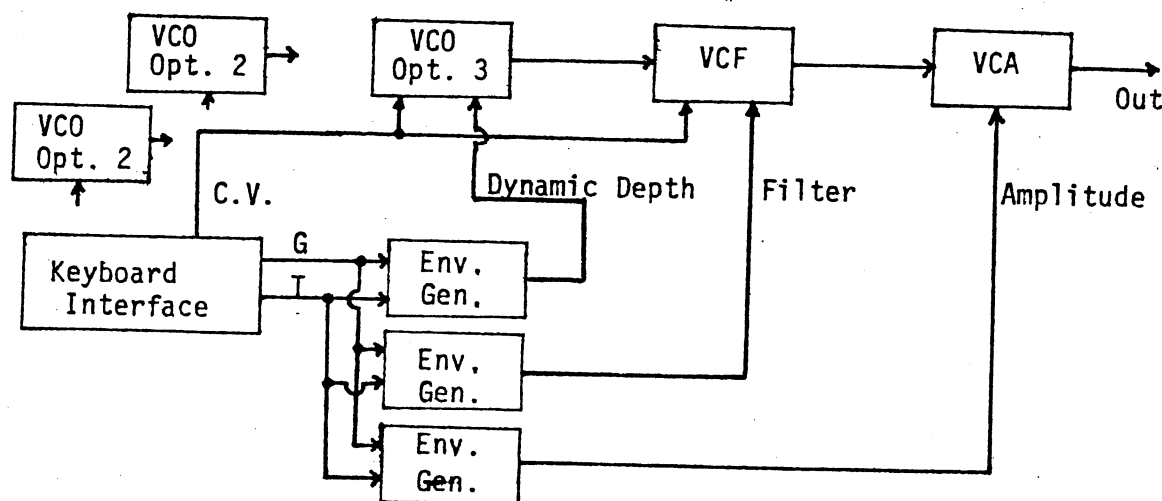
ENS-76 (option 2) [two]

Four Pole SSM-2040

ENS-76

ENS-76 Option 2

The patching of these modules that we have in mind is as indicated below:



In the patch above, we do not use two of the VCO's, but you can think of them as being in the patch in parallel or to provide vibrato, etc. Study of the patch will show that it is somewhat standard except for the inclusion of a third envelope generator to control the depth of self-modulation (or external modulation using one of the unused VCO's). This gives us an extra "handle" on the timbral variations and is the reason we have selected the ENS-76 Option 3 as the primary VCO for the system. Another more standard approach is also easily available to us with this combination of modules. We would use one of the VCO Option 2 modules as the primary oscillator, and then use the third envelope generator to vary the resonance (the Q) of the filter. We chose the four-pole filter for this setup because it has traditionally been found to be more useful for imitative synthesis. The ENS-76 option 2 envelope generators were selected instead of the SSM-2050 design because we wanted the delay feature which is often useful where as many as three envelopes are involved. Assuming a cost of \$100 for the keyboard itself, and another \$100 for a suitable cabinet and panel for the synthesizer, about 3/4 of the initial funds will be exhausted. We can perhaps find as much as \$100 for extra devices, of which a balanced modulator may be one of the first choices. This device will be useful if there is a singer in the band. It is a little strange that we would want to produce a sound as wierd as a ring modulated voice when we have gone to the trouble of producing an imitative system. However, this is often done, probably for its attention drawing effect.

As in the above scenarios, the use of the remaining expected 1 year funds will depend in part on experience with the devices built so far. There is an additional problem here however in that we may have used all the available panel space long ago,

since performance oriented synthesizers tend to be as small and compact as possible. There should still be spaces for adding such things as noise sources, low-frequency oscillators, and sample-and-hold units, and in your initial planning, it is well to plan space for these. However, it will probably be impossible to add any units that require a great amount of panel space. For this reason, it is not a bad idea to consider building a separate "accessory unit" or "wing" for your synthesizer. Central to this unit will probably be one or two major units such as a frequency shifter, analog delay line, sequencer, etc. These devices will perhaps add new dimensions to the type of music you can produce with your synthesizer and band. At the same time you will be able to tuck into this wing the extra small devices you have been needing. The wing should be able to share the same power supply with the main unit, and also you will probably want to run over, or be able to run over, other signals from the main synthesizer.

#### SCENARIO 4: Expansion of an Existing Academic Studio

Here we assume that the builder is in charge of expanding an existing academic electronic music studio which already has a studio quality synthesizer in operation. The expansion may include the addition of some conventional modules, but it certainly includes the addition of new experimental modules. The following goals and resources are assumed:

|                                    |                                    |
|------------------------------------|------------------------------------|
| Musical Goals:                     | Academic and experimental          |
| Engineering Goals:                 | Support of music                   |
| Immediate Budget:                  | \$300                              |
| Expected additional funds (1 year) | \$300                              |
| Time Requirements:                 | New device added every month or so |

If the initial expansion includes the addition of traditional modules, then most of what was listed for Scenario 2 will apply here. It is in the area of new modules that things start to get more interesting. What we have in mind here is that the technician who maintains the equipment and assists the composers will spend his extra time building new devices and integrating them into the system. Thus we have assumed an initial \$300 and additional funds of \$300, which are to be used for parts. We now list some of the experimental devices which we have found useful which can be added to a conventional synthesizer:

- Variable Slope VCF from ENS-76 Series, EN#72
- Multi-Phase Waveform Animator from EN#87
- Timbre Modulator from ENS-76 Series, EN#72
- Timbre Modulator from EN#84
- Through-zero VCO (ENS-76 VCO Option 4 or from EN#65 by Jan Hall)
- Second-Order Envelope Generator from EN#86
- Frequency Shifter from EN#83
- Slew-Limited Sample-and-Hold from EN#61
- Digital Noise Source, ENS-76 Series, EN#76
- Analog Delay Line using MN3005, Jan Hall, EN#87

We estimate the components cost of the above modules at about \$230, so allowing for panel and cabinet provisions, this list could easily use up the initial \$300 and begin to work on the second \$300. This would also seem to work in nicely with the time requirements. Of course, there will always be new things coming along in the newsletter to keep this list going.

APPENDIX B

ACE Electronic Parts 2  
5400 Mitchelldale B-8  
Houston, TX 77092

Active Electronics 1a  
PO Box 1035  
Framingham, MA 01701

ADVA 1  
Box 4181  
Woodside, CA 94062

Advanced Computer Prod. 4  
PO Box 17329  
Irvine, CA 92713

Aldelco 2  
2281 Babylon Turnpike  
Merrick, NY 11566

ALF Products 4  
128 South Taft  
Denver, CO 80228

Altaj 1  
PO Box 38544R  
Dallas, TX 75238

Analog Devices Inc 3b  
PO Box 280  
Norwood, MA 02062

Ancrona Corp 1  
PO Box 2208  
Culver City, CA 90230

Babylon Electronics 1  
PO Box 41778  
Sacramento, CA 95841

Brigar Electronics 1  
10 Alice St  
Binghamton, NY 13904

Burstein-Applebee 1c  
199 Mercier  
Kansas City, MO 64111

CFR Associates 1d  
Box F  
Newton, NH 03858

Chaney Electronics 1  
PO Box 27038  
Denver, CO 80227

Circuit Specialists 1  
Box 3047  
Scottsdale, AZ 85272

Components Center 1e  
Box 295  
West Islip, NY 11795

Computalker Consultants 4  
PO Box 1951  
Santa Monica, CA 90406

Cita Electronics 1f  
Oakland St PO Box 2  
Needham, MA 01913

Devtronix Organ Products 3g  
5872 Amapola Dr  
San Jose, CA 95129

Diamondback Electronics 2  
PO Box 194  
Spring Valley, IL 61362

Digi-Key 1o  
PO Box 677  
Duluth River Falls, MN 56701

Digital Research Corp. 2  
PO Box 401247  
Garland, TX 75040

EDI 1p  
4900 N. Elston St  
Chicago, IL 60630

Ellie Elect. Inc. 1  
700 Hempstead Tpke.  
Levittown, NY 11756

Electronic Discount Sales 2  
138 N. 81st St  
Mesa, AZ 85207

Electrovalue 1h  
PO Box 337  
Peterborough, NH 03458

Eltron 2  
PO Box 2542A  
Funnyvale, CA 94087

Fair Radio Sales 1i  
1016 E. Eureka Box 1105  
Lima, OH 45802

Formula International Inc 2  
12603 Crenshaw Blvd  
Hawthorne, CA 90250

Reichert Sales 2  
10 E. Garvey Ave  
Covina, CA 91790

Godbout Electronics 1r  
Box 2355  
Oakland Airport, CA 94614

Herbach & Rademan 1  
401 E. Erie Ave  
Philadelphia, PA 19134

Integrated Ckts. Unlimited 2  
7889 Clairemont Mesa Blvd  
San Diego, CA 92111

International Components 2  
1208 Bowling St  
Columbia, MO 65201

International Elect. Unl. 1  
Village Sq., PO Box 449  
Carmel Valley, CA 93924

Ithaca Audio 4  
PO Box 91  
Ithaca, NY 14850

Jade Computer Products 2  
5351 W. 144th St  
Lawndale, CA 90260

James Electronics 1q  
1021-A Howard Ave  
San Carlos, CA 94070

J. B. Saunders Co 2  
3050 Valmont Rd  
Boulder, CO 80301

Meshna 1j  
PO Box 62  
E. Lynn, MA 01904

New-Tone Electronics 1k  
PO Box 1738A  
Bloomfield, NJ 07003

Page Digital Electronics 2  
135 E. Chestnut St 4A  
Monrovia, CA 91016

Quest Electronics 2  
PO Box 4430E  
Santa Clara, CA 95054

Ramsey Electronics 2  
Box 4072B  
Rochester, NY 14610

Radio Hut 2  
PO Box 38323  
Dallas, TX 75238

S.D. Sales Co 1  
PO Box 28810  
Dallas, TX 75228

Solid State Inc 1k  
46 Farrand St  
Bloomfield, NJ 07003

Solid State Music 4  
2102A Walsh Ave.  
Santa Clara, CA 95050

Solid State Sales 1l  
PO Box 74D  
Somerville, MA 02143

Star-Tronics 1m  
PO Box 683  
McMinnville, OR 97128

Tel Labs 3n  
PO Box 375 - 154 Harvey Rd  
Londonderry, NH 03053

The Digital Group 4  
PO Box 6528  
Denver, CO 80206

Tri-Tek Inc. 1  
6522 N. 43rd Ave  
Glendale, AZ 85301

VMP Electronics Mart 2  
PO Box 415  
Carmel Valley, CA 93924

Curtis Electromusic  
110 Highland Ave  
Los Gatos, CA 95030

Sound Logic  
PO Box 49331  
Austin, TX 78765

Mid America  
2309 South Archer Ave  
Chicago, IL 60616

Reticon Corp.  
910 Benicia Ave  
Sunnyvale, CA 94086

Don't bet that very many  
of these are still around!

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## ELECTRONOTES

APPLICATION NOTE NO. 131

1 PHEASANT LANE

May 4, 1979

ITHACA, NY 14850

TROUBLESHOOTING - 1

(607)-273-8030

Troubleshooting is something that nearly everyone involved in electronics must do for a certain percentage of his time. Some of these people (called rapairmen) do it all the time and have a high degree of skill, at least in troubleshooting certain types of systems. Others do very little troubleshooting, or at least they probably think they do very little. In this note, and the next three notes to follow, we will be looking at the practice of troubleshooting. We will try to dispell some myths about it, and point out how lack of experience can be compensated by related experience, and by just ordinary reasoning.

First, we want to point out that there is a difference between troubleshooting and repairing. Troubleshooting is really "troublefinding" and repairing is then a simple matter of correcting a problem that has been identified. Troubleshooting is an activity that will take an undetermined amount of time. Time to repair is usually well determined. Troubleshooting is not something that can be taught in its most general sense. Repairing can be taught. Persons taught to "troubleshoot" a single system (say a certain brand of TV) are really repair technicians. Since standard learned problems will be the actual situation encountered in 99% of their cases, the problem identification is fast and automatic, and their main activity is repair. In these notes, we want to discuss a more general type of troubleshooting, where many different systems are presented, each to be studied and handled.

The second preliminary point to make is that since we are concerned with a very general sort of troubleshooting, we can benefit greatly from troubleshooting experience well outside the area of electronics. This is why we say above that some persons believe they do not have such experience, but actually, they do. The person who identifies problems with his car is a potentially good electronics troubleshooter. The person who identifies problems with plants growing in his garden is a potentially good electronics troubleshooter. Any person good at locating problems in his own particular area will be good in locating problems in other areas, provided he does transfer his methods after identifying them.

The third preliminary remark is that a person given to rational thinking, and who is otherwise "good at getting things done" will be a good troubleshooter. Such a person will think out the situation, and will have the resourcefulness to make the necessary tests to find the answer.

One point of these preliminary remarks is to make it clear that to the extent that you do not adopt these procedures, points of view, and "go getting" attitudes, you will have problems troubleshooting. The relation between troubleshooting types of thinking and general thinking is reciprocal. Becoming a good troubleshooter may well imporve your general thinking, and psychologically improve your outlook toward your ability to handle problems in general.

STEPS IN TROUBLESHOOTING:

Once you have done some troubleshooting, you just do things without thinking much about the actual steps involved. Here we want to say some things about what some logical steps are.

The first step is to determine exactly what is wrong. If a unit is completely

dead, than that is exactly what is wrong with it, but for semi-functional units, you have to obtain more information. If a multi-waveform oscillator has no sine wave output for example, are the other waveforms also missing? If a unit drifts, how fast does it drift, and does it seem to drift with temperature? If a unit is inaccurate, is it 1% off or 50% off? These are all important questions. As a rule, get enough information about the problem situation so that you could describe it to another person. Have the answers to the questions this other (imaginary?) person will ask.

It is such a natural reaction to monkey with every available knob when something goes wrong, that often a unit with a problem would not work properly even if the original problem were to disappear by itself. Thus, in checking out a problem with a unit, it is useful to note all knob positions, without making any further changes. Often you will find an improper setting and suppose with great relief that you have located the problem with ease. Of course, correcting this incidental condition will not solve the actual problem. Reset controls to the proper positions, but make sure you have some record of the actual settings when the unit first was declared in need of repair.

Before you do anything at all toward actual troubleshooting (like removing a unit from a complex housing), THINK. Think about what could be wrong. Do not do anything other than very simple visual inspection or make simple supply voltage checks without having something in mind about what could be wrong, and what you are checking.

Of course, the classical technique of troubleshooting is to isolate the problem. This is necessary and it does work, but is not always simple. If electronic circuits all had a chain structure, or a branch structure, so that you could always check from the output back, then isolation would be simple. However, modern electronics involves itself with loops of many sorts. When you have a loop, failure in one portion may well, and usually will, cause the whole loop to appear to be malfunctioning. Thus you will not be able to isolate by circuit malfunction. In such a case, you must isolate by component malfunction. This means that you do not worry about what the whole circuit is supposed to be doing, but rather consider each component (transistor, op-amp, etc.) to see if it is behaving consistent with its basic operation and its input environment. In short, you check each component to see if it is doing "the best it can" under the circumstances even if what it is doing is nothing like what it would be doing if the circuit were functioning properly. Find a component that is not playing by its own rules and probably you have isolated the fault in the loop and at the same time, found the component to change to affect repair.

Once you believe you have found the problem section, you will not be able to resist the temptation of pulling a suspected component and replacing it. Don't try to resist the temptation, but do slow down a bit. Make sure you have a good enough record to put things back the way they were when you find out that you were wrong, as you often will. Remove components carefully, because you may well have to put them back. Also, if the part to be replaced is expensive and/or hard to get, you should not attempt to replace it without determining if there is some reason why the part failed. Perhaps there is a short, or another component that failed first, causing failure in the component you discovered. What's worse than a blown expensive component? Blowing the replacement, that's what.

If you follow these general steps, you will be able to repair perhaps 50% to 75% of the units you encounter. It is the remaining units which will eventually make an expert out of you. How do you find what is wrong when nothing is wrong (nothing seems to be wrong - that is)? How do you find the problem when it is intermittent (sometimes there, but not always; and usually, not often)? The next few notes will give some more ideas and suggestions.

1 PHEASANT LANE

May 11, 1979

ITHACA, NY 14850

TROUBLESHOOTING - 2

(607)-273-8030

## GOLDEN RULES OF TROUBLESHOOTING

The following Golden Rules present ideas that are important to troubleshooting. Read them now, try to remember them, but most importantly, when you are about to give up, come back and check out these rules one by one.

G.R. No. 1 "You Probably Screwed It Up When You Did \_\_\_\_\_"

Given a subsystem B of a larger system A, a failure in B that follows immediately a change in A, no matter how unlikely the link seems, is linked to the change in A.

Let me begin with an actual example, a radio receiver A with a subsystem B consisting of an audio section. I installed an output jack in A. After putting the radio back together, the sound from the speaker was distorted, even with nothing plugged into the jack I had installed. Even disconnecting the jack completely did not correct the problem. Conclusion: Something has coincidentally gone wrong with the audio section B. Wrong! G.R. No. 1 says in effect that there are no such coincidences. So where did I mess it up? To make a long story short, when the chassis was removed from the cabinet, one of the securing lockwashers fell and was attracted by the speaker magnet and remained in the speaker. This caused the distorted sound. This sort of problem is very difficult to locate - but I would have saved a lot of time if I had not tried to find a problem in the audio

Convincing yourself that the link is there, believing G.R. No. 1, will not always help you find the actual problem, but it will tell you where the problem is not located.

G.R. No. 2 "The System May Ignore The Designer, But It May Not Ignore Nature"

Every system with unsatisfactory performance is still doing exactly what it is supposed to under the circumstances. It is still obeying the laws of nature even though it is not obeying the laws of the designer.

G.R. No. 3 "The Impossible Does Not Happen Even Though It May Appear To Be"

Many things are impossible in engineering. Thus, the appearance of something that appears to be happening, and is yet known to be impossible, is a strong indication of a malfunction to be investigated.

What G.R. No. 2 and G.R. No. 3 say is that you have to believe your own eyes. Wishing, or supposing that things were otherwise than they are will do no good, and saying that something cannot be happening because it is impossible will not stop it from happening. You must realize that a malfunctioning system is still a system, and it will usually do something. Thank goodness it does something - it is very difficult to find a problem in a completely dead circuit. The system is giving you information, telling you how it responds to the laws of nature. It is not laughing at you! Use the information it gives you, every bit of it. The strangest thing that appears to be happening may be your best clue as to what is wrong, even though it often will not point directly to the thing that is wrong.

G.R. No. 4 "When You Know Something You Know It, But . . . ."

When testing a malfunctioning system, the measurements you make will tell you things that you then must believe. Believe them as long as possible, but not forever (if you can't solve the problem). Know how long to hold on, and when to give up.

This rule is best illustrated with an example. Suppose you have checked a system with a scope and find an unwanted oscillation at one stage. You know it is there because you saw it on the scope. You conclude that this is causing the problem, and proceed to study the problem further. You can find no other evidence that the oscillation is present, and you can not solve the original problem. After considerable study and frustration, you find that it was the capacitance of the scope probe that was causing the oscillation - it was not present when the scope was not connected - and further, the actual problem with the circuit had nothing to do with the oscillation. This example also illustrates the following Golden Rule:

G.R. No. 5 "Just Finding Something Wrong Is Not Necessarily Finding The Problem"

A system may have a series of malfunctions, x, y, and z for example, occurring in order. Assume x and y are minor, and do not call attention to themselves, but z is more serious, and must be fixed. Locating and correcting x and/or y will not correct fault z, even though they may lessen the seriousness of z.

This sort of situation is not uncommon, and can be very frustrating, first because you did find and correct one problem, and secondly because it is not unusual for the correction of a minor problem to cause a slight improvement in the major problem, causing you to believe you are on the right track while actually you are still in the wrong area.

G.R. No. 6 "Most Faults Are Simple Faults"

A malfunction of a system is usually due to the malfunction of a small component part, a relatively simple fault. This simple fault may be due however to a complex causal chain of events.

What is being said here is that while a fault may occur through a complex chain of events, the actual fault will be simple. For example, a transistor may be blown out, but it won't likely have suddenly gone non-linear, or decided to reject a certain frequency. This means that at times it is better to give up the "how" and worry more about finding the actual "what." It is always nice to know exactly how something went wrong, but in many cases, you will have to give up and resort to a more "trial and error" procedure, the extreme case of which is the procedure of changing every component in turn. This will work if you have isolated the fault to one system and are sure it is not caused by some external factor.

G.R. No. 7 "Don't Work Too Steady On Troubleshooting"

If you work at troubleshooting a system for an extended period of time without any success, take a break. Do something else. This will break up the well worn ruts in your thinking and clear out mental cobwebs. Probably a very difficult problem will be most efficiently attacked by about 1/4 work, 1/4 thinking, and 1/2 doing something else.

G.R. No. 8 "When Working Against A Deadline, Be Extra Efficient, Not Less Efficient"

With a deadline coming, it is all too easy to flail away at this and that and end up with nothing. When time is limited, you must be more efficient. Do take the time to take adequate (but not elaborate) notes so that you can avoid doing things twice.

THE GOLDEN GOLDEN RULE: "If You Don't Know How It Is Supposed To Work, Then You Can't Expect To Find Something That's Not Right!"

1 PHEASANT LANE

May 18, 1979

ITHACA, NY 14850

TROUBLESHOOTING - 3

(607)-273-8030

WHAT CAN GO WRONG WITH CIRCUITS?

What can go wrong with circuits? Yes, just about anything, but usually there are things that are very unlikely to go wrong under certain circumstances. Here we will begin by listing certain circuit situations, and the listing is in order of least likely to have something wrong, to most likely:

Least Likely



Most Likely

1. Commercial Circuit Without Stress
2. Commercial Circuit With Stress
3. Homemade Circuit That Once Worked
4. Homemade Circuit Under Construction
5. Homemade Design Under Construction

We should define a few things. Stress is something like physical motion or internal or external heating. A homemade circuit is one you build, but which was designed for home builders. A homemade design is something you designed and are now trying to build. Note that there is a large difference between these circuit situations, and while general troubleshooting rules apply, you will save a lot of trouble if you consider first your situation. Of course, the listing above is only "in general" and there are always exceptions.

COMMERCIAL UNITS: We say that there is very little that can go wrong with a well made commercial unit - unless it is subjected to stress. Perhaps it is better to say that if something does not go wrong soon after you get the unit (early failure) and if there is no stress on the unit, very likely nothing will go wrong. Put another way, if you have had a working commercial unit that has gone out, it is probably due to some sort of stress.

What is stress and how do you look for it? First you look for it. Are there any loose wires, any burned out components, any dents or foreign objects? If there is no sign of such obvious stress, look for hidden stress. Where is heat generated in the unit? Where are plugs inserted and removed frequently? What knob controls or switches are of inferior quality and are likely to have been overused? Keep in mind that stress is applied or internal. Lightning may enter the unit through the power lines (applied) or the heat of the power transistors (internal) may have burned themselves out. Always look first for stress caused by heating and cooling of a portion of the unit. Next look for physical abuse. Finally, since stress from an external source must be applied, where are the entry points? What input line might have come in contact with high voltage? What output line may have become shorted?

What about a component just going bad? We seem to have ignored that possibility. It can happen, but don't look there first. You will find ten worn out switches for every burned out IC. It is the nature of modern solid state circuits to just keep working. Mechanical things on the other hand still wear out as before, and people still tend to abuse their equipment as before.

HOMEMADE UNITS: When there is a problem with a homemade unit, it makes a very big

difference whether or not the unit ever was completed and found to work successfully. If it ever did work properly, then the situation is thrown more toward that of the commercial case above, except you will look first at areas of construction failure - wires that were not soldered properly, etc.

Far more common, and we suspect particularly common among our readers who are for the most part builders, will be the case where something built at home never has worked. We can assume that you have built your unit from a set of plans, and that the plans were correct (always check to be sure that corrections were not published later!). In all such cases, if you get good parts, follow instructions carefully, and build carefully, things should turn out well. Getting good parts is largely a matter of not buying obvious junk (such as 10 for the normal price of one, or "U-Test-Em" parts). Very few newly built circuits will fail to work because of a bad part. It is much more likely that instructions have not been followed (wiring errors), or that there is an area of poor construction. With the small size of the circuitry, it is all too easy to miss a connection, have something that appears soldered missed, or to have a blob of solder shorting two or more points. Thus, you should check first the correctness and the quality of the wiring job.

To check the correctness, you should first study the schematic carefully to make sure you understand what the interconnections are supposed to be. Then, make a photocopy of the schematic. Now, check the wiring and cross out sections you find to be correct (a colored marker is good for this). The error, or missing component may suddenly appear.

With the correctness verified, check the quality. Do not, as some suggest, go through and reheat every solder connection. This may create more errors rather than remove existing errors. You can, if the circuit is spread out, reheat (and add a bit more solder) connections that are in the clear. However, connections that are easy to get at are also those that are most easy to inspect visually. Play the odds instead. You are most likely to have made a bad connection in a spot that is difficult to reach. It is because it is difficult to reach that you did a poor job in the first place, and further, this tightness will make it more difficult to find and correct the problem. It may be useful in such a case to check the circuit under power. Use a small probe such as a length of hookup wire with about 1/8" of insulation removed. Check for the proper signal on the pins of the IC. If it is not there, or if you are unsure what the signal should be, check to make sure you have the same signal on the pin and on the conductor leading up to the pin. If it is different, you have found your bad connection. If it is really difficult to get at, remove some components to make things easier, and then put them back later. It's easier in the long run that way.

In some cases you will be sure that some section is bad. It may be a connection or a bad part. Perhaps the easiest thing to do is to take apart and rebuild that section. Typically, you will remove an op-amp or two (save to be retested later, but replace them now) and assorted other parts. I do this frequently, and it usually works. Later a retest of the parts removed shows about 3/4 were good. Thus the cause was a bad connection which was difficult to find.

In the event that your circuit is correct and properly constructed, and still does not work, then you troubleshoot it as a commercial unit, keeping in mind of course, that you still may have overlooked a construction error.

HOMEMADE DESIGNS: This is the area where you have designed your own circuit, and anywhere from 0% of it to 100% of it may be considered original. Very little can be said here except that general troubleshooting rules do apply. This is an area where more than ever you will be saying "It just has to work" despite the fact that it does not. Perhaps the only general thing to say is that it pays to keep a somewhat more general view of what you are doing. Don't become bogged down on a small subsystem because such a small subunit may not work because it is not being treated properly by the rest of the circuit. Keep the big picture in mind, and make sure things work as you really think they do.

1 PHEASANT LANE

May 25, 1979

ITHACA, NY 14850

TROUBLESHOOTING - 4

(607)-273-8030

PRACTICAL CIRCUIT TROUBLESHOOTING

There comes a time when theory ends and troubleshooting proper must begin. You have your circuit on the bench, have reviewed possible causes of the trouble, (and are "sure" you know exactly what is going to be found wrong!), and are ready to complete the task. You may be 95% finished with the repair at this point, or you may be just starting.

EQUIPMENT:

I know of no single instrument more important than an oscilloscope. In fact, you can measure nearly everything you need to know with a scope. The scope is your eye for looking into the circuit. If you have other equipment, that's just fine, a simple ohmmeter or a digital meter, and other things like signal generators and frequency counters will be handy at times.

THE ATTACK:

There is no reason not to attack immediately the one cause you suspect the most. Don't even check for the power supply levels first. Go right to the point you suspect. The one time in ten that you get it right just by thinking will more than make up for long hours of frustration. If this does not work, go on to use the isolation techniques and other techniques we have been discussing.

THE BAD AREA:

Eventually you will find an area of trouble. This will be characterized by a voltage or signal that is not what it should be. It is important to realize that you may have found an effect rather than a cause. For example, there might be a case where you are expecting a DC zeroed AC signal at the output of an op-amp, and instead you find the positive supply voltage (output "pinned"). The next thing to check is to see if the differential input voltage is zero or positive. Chances are that it will not be zero. Probably it will be some positive value (the + input will be more positive than the - input). Now, even though the output of this op-amp is the wrong value, you have no reason to suspect that there is anything wrong with this op-amp, because the response you see is exactly what you would have if the op-amp were perfectly good. The op-amp is not receiving the correct signal through its input structures, so it can't possibly be doing the correct thing. You have located an effect, not a cause.

In general, don't expect trouble with any IC, and expect it even less if an IC is obeying its own rules. This applies to op-amps and to digital logic IC's. Some components are difficult to evaluate because they don't have any rules except when they are operating in their correct modes, so you can't be sure with these. Trace the trouble area backward until you find an IC that is not obeying its own rules, or until the problem ends. Most likely, you won't find a bad IC, but rather the problem will disappear up the line. Now look exactly in this area to find out why the closest IC is not getting the signals it is supposed to.

TO CHANGE A PART OR NOT TO CHANGE:

During the tube days of television, about 9 out of 10 problems with TV sets

were due to bad tubes. Thus the repairman would look at the screen and decide which section seemed to have a problem and then start substituting in new tubes until the problem went away. However, with solid state circuitry the active components have become as reliable or more so than the passive ones, and are somewhat more reliable than any mechanical ones in general. Thus, part switching is not too useful. Whether it is worth your time or not depends on how much trouble it is to remove it, and how much risk there is that some other fault in the circuit may blow the new part just the way it did the old one. Certainly in a breadboard situation where you are working on a new circuit it is easier to change to a new part than it is in a commercial unit where an IC is soldered in and hard to reach. Also, many digital circuits are both built on boards using sockets and relatively difficult to troubleshoot, and here it may be useful to change IC's, particularly if the board did work at one time. Note that bad IC's are a much more common cause of trouble in digital circuits. This is not because the digital IC's are less reliable, but more likely because there are many more IC's in your average digital circuit, and because most digital circuits are pretty much all IC's and conducting paths.

#### BRINGING OUT TEST SIGNAL LINES:

The usual type of scope probe is well suited to grabbing on to something like a wire lead, but grabbing on to the foil side of a PC board of a commercial unit is something else. You are lucky if you can just locate the point to be tested and then touch the probe to that point. However, it is sometimes just about necessary to have a permanent clamp to a test point. Even if you do manage to get a grip, the probe may fall off, or worse still, slide and short together two PC traces. In such cases, it is useful to tack solder a length of hookup wire to the appropriate point, run the other end to a safe out-of-the-way place, and clip the scope probe there. You can leave the hookup wire attached until you no longer need the test point.

Another useful trick is to keep in mind that most traces run from some point to some other point. You may be trying to clip on to a difficult point while a perfectly accessible point is available right in front of you. Don't waste time trying to clip to a blob of solder when the other end of the conducting path is a readily available pot terminal or panel switch.

#### REPLACING PARTS KNOWN TO BE BAD:

Suppose you have found a bad part, and have a replacement in your hand. It should be a simple matter to change it, and sometimes it is. However, it is also relatively easy to mess other things up in the changing process.

Even something as simple as changing a socketed IC can be a problem. There are special tools for prying the IC's out of their sockets, and while you don't need these, you do need some sort of tool. If you try to lift the IC out with your fingers, somewhere in your career, one of these bugs will jump out, flip over, and usually end up with its feet digging through the skin of your thumb. While it is easy to remove the IC from your thumb, it is not all that pleasant an event. Instead, pry it up, lifting uniformly with a small screwdriver or a knife blade. Wait until it is out before picking it up with your fingers. It is hard to believe, but those bugs do bite!

The main problem in removing a soldered in IC is that you won't be able to heat all the leads and pry out at the same time. Mini-DIP IC's are not so difficult, but 14, 16, and more pins present a substantial problem. To do these right, you need a thing called a "solder sucker." It is a little plunger operated vacuum cleaner with a teflon spout. You push in the plunger, heat the solder, immediately place the spout over the hole, and release the plunger. It sucks the liquid solder out of the hole and up the spout. It works, but costs money too. You can also clip all the IC leads and remove the pins one at a time if this is more convenient.



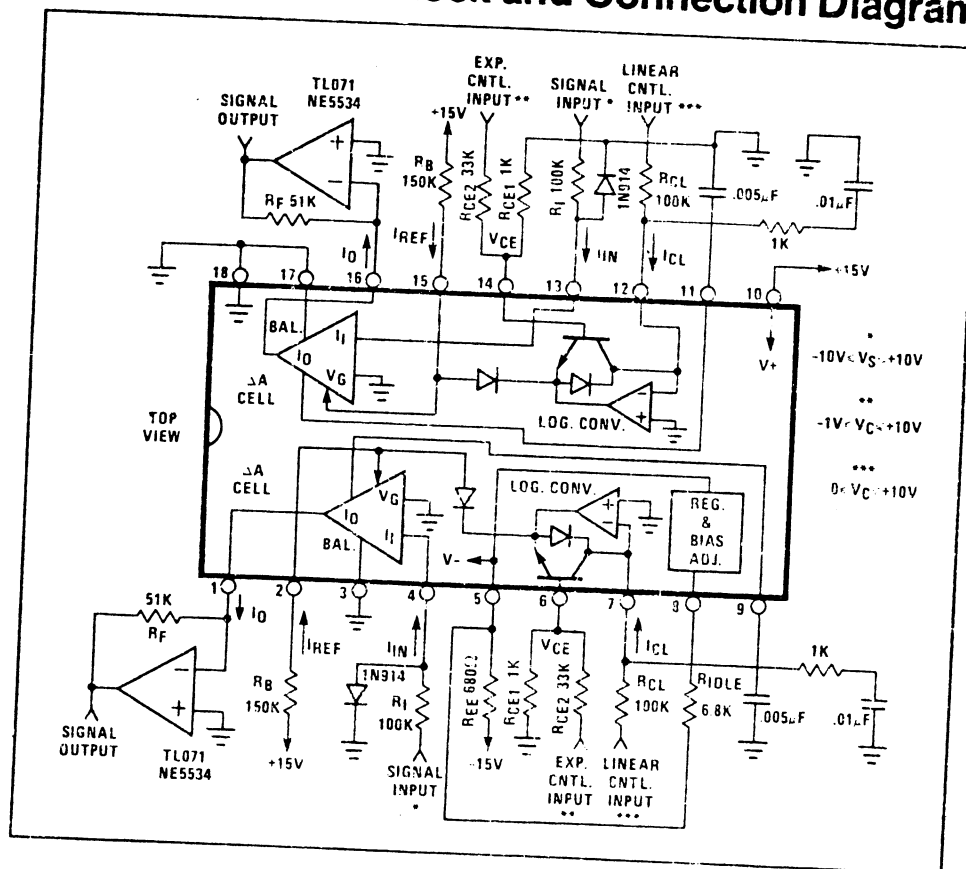
It is often the relative happy lot of the musical engineer that he designs something only to have it become available as a self-contained IC, or that he will find a new and better IC available for his designs. The unhappy part is that it is often difficult to keep design work up with available IC's. It is the purpose of this appendix to at least list the features of some of these new and special purpose electronic music IC's. Eventually, many of these will find their way into the main body of the Preferred Circuits section.

Many of the IC's listed here are made by either SSM (Solid State Micro Technology for Music, 2076B Walsh Ave., Santa Clara, CA 95050) or by CES (Curtis Electromusic Specialties, 110 Highland Ave., Los Gatos, CA 95030) who have complete data sheets available. The type 13600 OTA is available from National and from Exar, and the CA3280 is from RCA.

We will list these devices in order of VCA's, VCO's, VCF's and envelope generators:

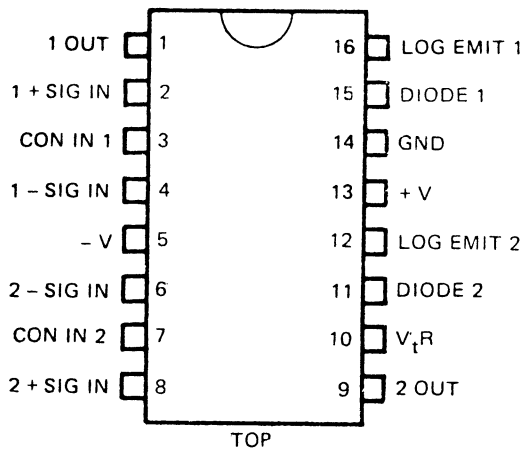
#### VCA's:

### CEM 3330 Circuit Block and Connection Diagram



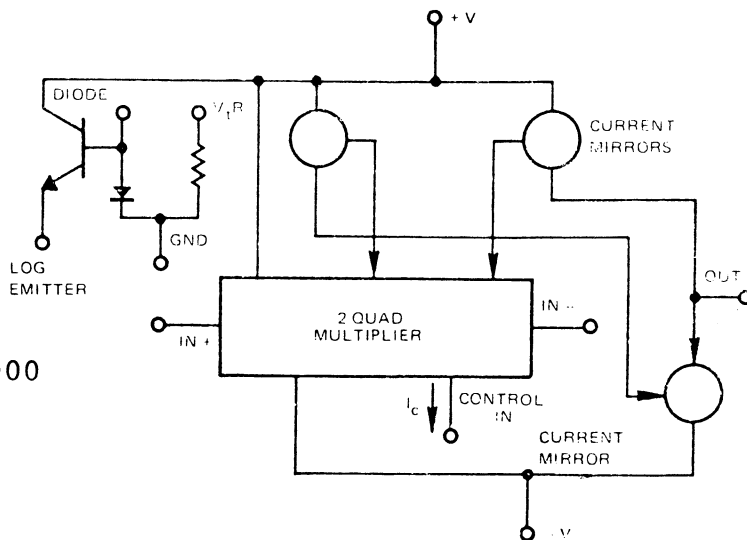
#### Features

- Low Cost
- Two Independent Voltage Controlled Amplifiers in a Single Package
- Simultaneous Linear and Exponential Control Inputs
- Wide Control Range: 120dB min.
- Very Accurate Control Scales for Excellent Gain Tracking
- Exceptionally Low Control Voltage Feedthrough: -60dB minimum without trim, better than -80dB with trim
- Low Distortion: Less than 0.1%
- Exceptionally Low Noise: Better than -100dB
- Class B to Class A Operation
- Summing Signal and Linear Control Inputs
- Current Outputs for Ease of Use in Voltage Controlled 2-Pole Filters
- Can Be Used in VCO and VCF Control Paths Without Causing Shift
- ±15 Volt Supplies

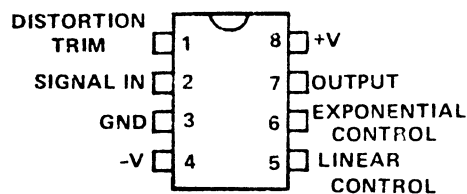


Pin Diagram

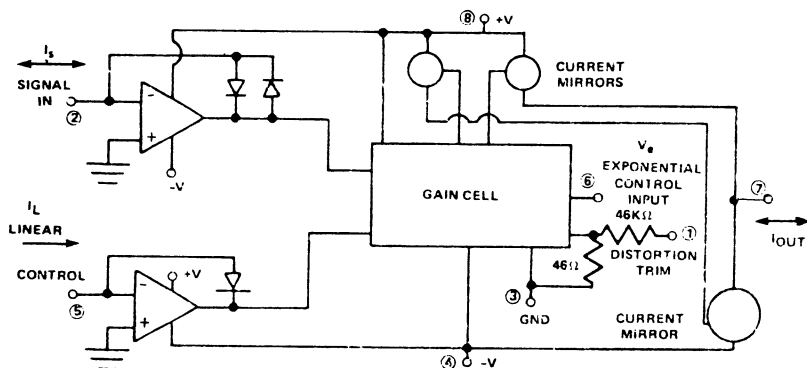
SSM2000



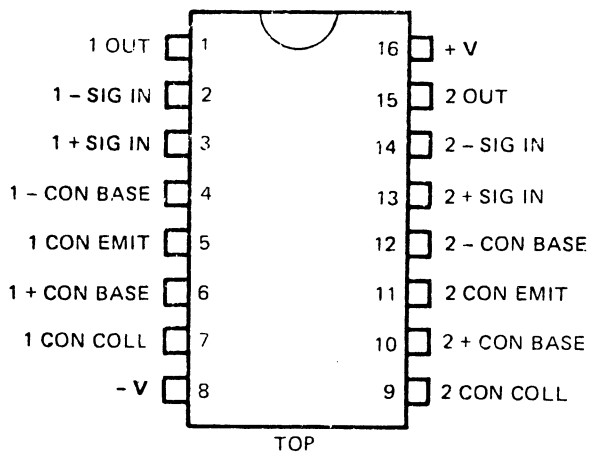
Equivalent Schematic (One Side)



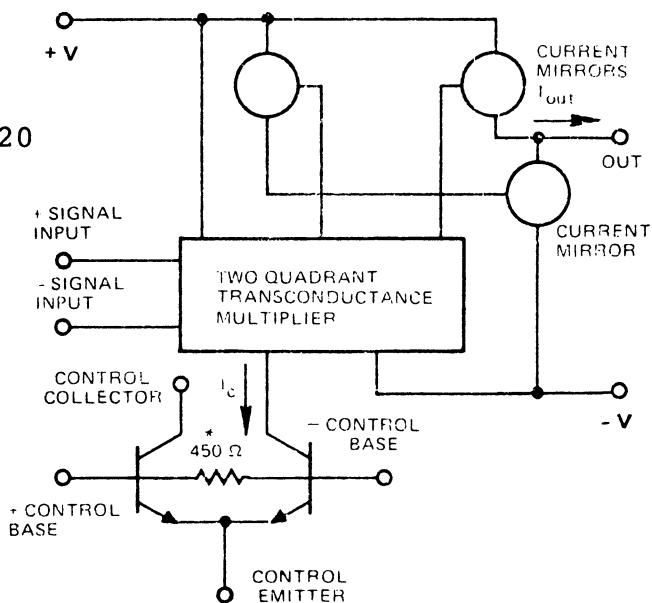
SSM2010



SSM2020

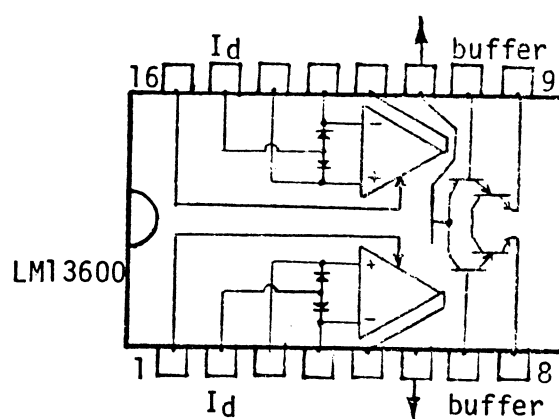
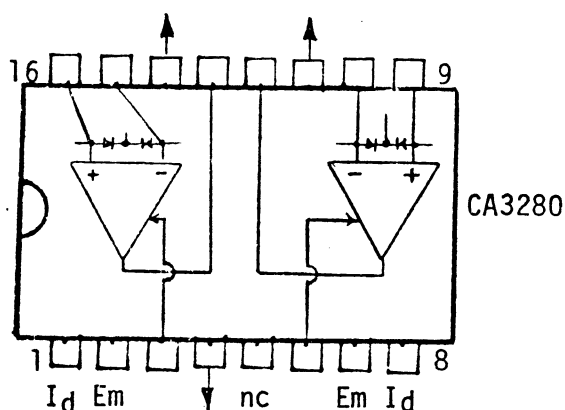


Pin Diagram



Equivalent Schematic (One Side)

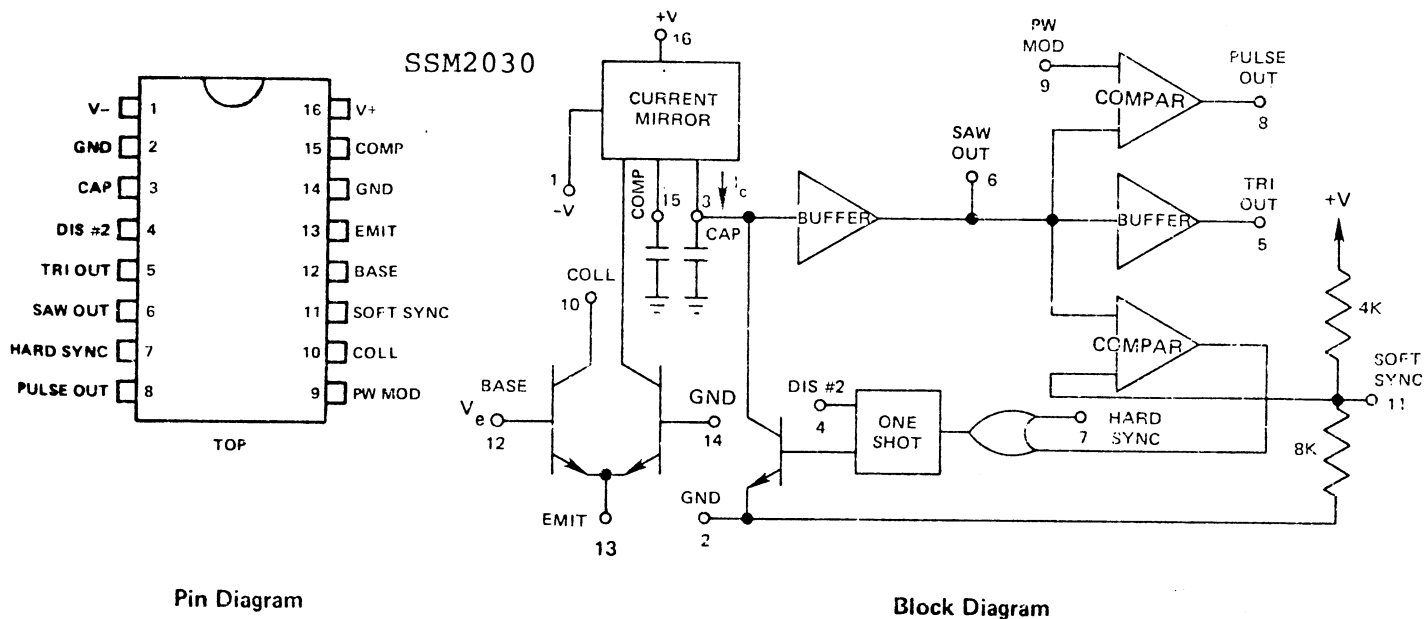
| Pin No. | CA3280                           | CA3080<br>equiv. | LM13600                          | CA3080<br>equiv. |
|---------|----------------------------------|------------------|----------------------------------|------------------|
| 1       | Amp. 1, Linearization $I_d$      | none             | Amp. 1, Cntrl. Current $I_{ABC}$ | 5                |
| 2       | Amp. 1, Tied Emitters            | none             | Amp. 1, Linearization $I_d$      | none             |
| 3       | Amp. 1, Cntrl. Current $I_{ABC}$ | 5                | Amp. 1, (+) Input                | 3                |
| 4       | Negative Supply (-15)            | 4                | Amp. 1, (-) Input                | 2                |
| 5       | N.C.                             | ----             | Amp. 1, Output $I_{out}$         | 6                |
| 6       | Amp. 2, Cntrl. Current $I_{ABC}$ | 5                | Negative Supply (-15)            | 4                |
| 7       | Amp. 2, Tied Emitters            | none             | Buffer 1, Input                  | none             |
| 8       | Amp. 2, Linearization $I_d$      | none             | Buffer 1, Output                 | none             |
| 9       | Amp. 2, (+) Input                | 3                | Buffer 2, Output                 | none             |
| 10      | Amp. 2, (-) Input                | 2                | Buffer 2, Input                  | none             |
| 11      | Amp. 2, Positive Supply (+15)    | 7                | Positive Supply (+15)            | 7                |
| 12      | Amp. 2, Output $I_{out}$         | 6                | Amp. 2, Output $I_{out}$         | 6                |
| 13      | Amp. 1, Output $I_{out}$         | 6                | Amp. 2, (-) Input                | 2                |
| 14      | Amp. 1, Positive Supply (+15)    | 7                | Amp. 2, (+) Input                | 3                |
| 15      | Amp. 1, (-) Input                | 2                | Amp. 2, Linearization $I_d$      | none             |
| 16      | Amp. 1, (+) Input                | 3                | Amp. 2, Cntrl. Current $I_{ABC}$ | 5                |



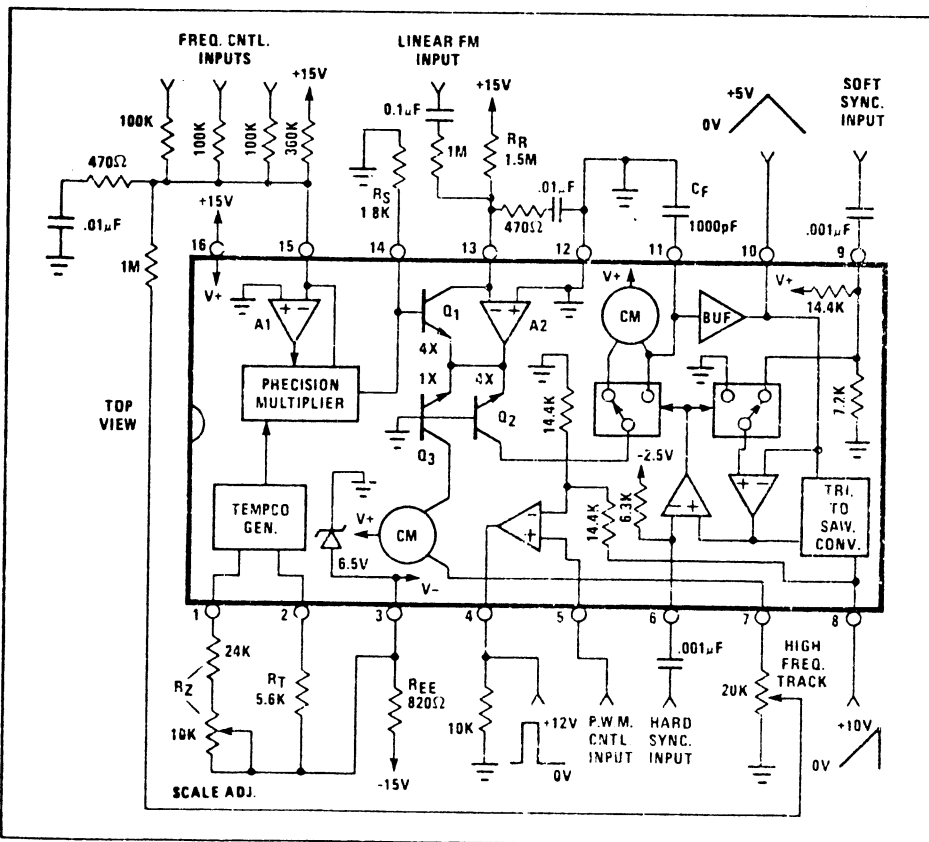
### ELECTRICAL CHARACTERISTICS

| Parameter                     | CA3280   | LM13600  | CA3080*  | Units        |
|-------------------------------|----------|----------|----------|--------------|
| Max. Supply                   | $\pm 18$ | $\pm 22$ | $\pm 18$ | volts        |
| Max. $I_{ABC}$                | 10       | 2        | 2        | ma           |
| Max. Differ.<br>input voltage | $\pm 5$  | $\pm 5$  | $\pm 5$  | volts        |
| Diode Bias Max                | 5        | 2        | n.a.     | ma           |
| Short Ckt. Protect            | $\infty$ | $\infty$ | $\infty$ |              |
| Bandwidth                     | 9        | 2        | 2        | MHz          |
| Slew Rate                     | 125      | 50       | 50       | v/ $\mu$ sec |

\*CA3080 included for comparison



## CEM 3340 Circuit Block and Connection Diagram

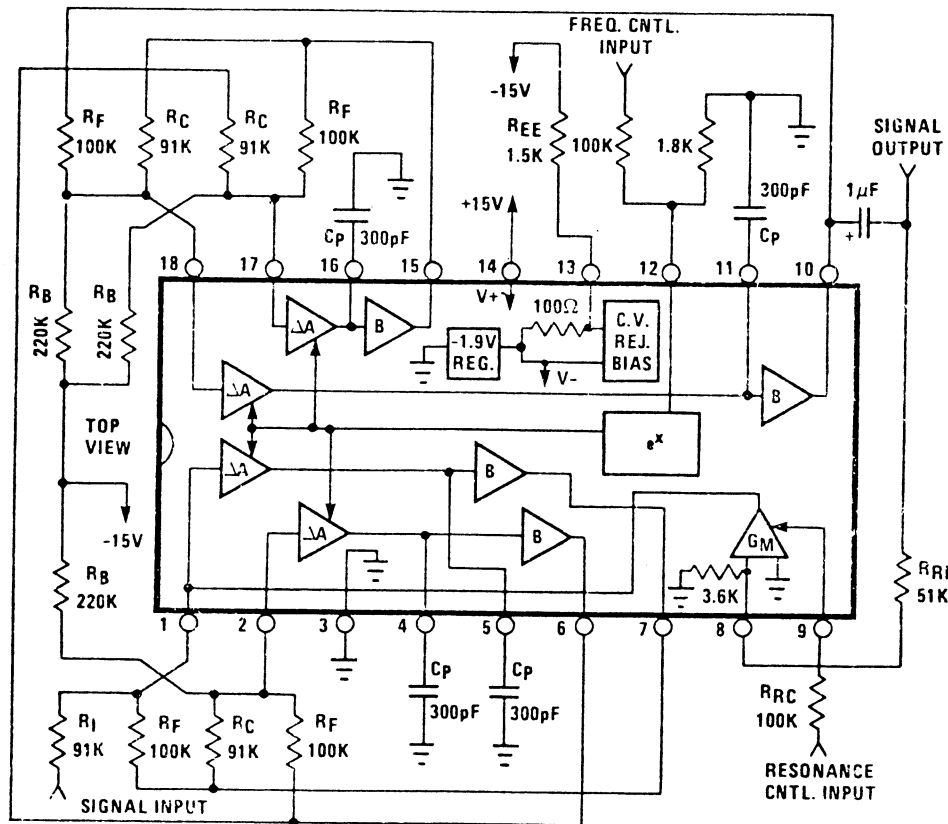


### Features

- Large Sweep Range: 50,000:1 min.
- Fully Temperature Compensated; No Q81 Resistor Required
- Four Output Waveforms Available; No waveform trimming required.
- Summing Node Inputs for Frequency Control
- High Exponential Scale Accuracy
- Low Temperature Drift
- Voltage Controlled Pulse Width
- Hard and Soft Sync Inputs
- Linear FM
- Buffered, Short Circuit Protected Outputs
- $\pm 15$  Volt Supplies

# Circuit Block and Connection Diagram

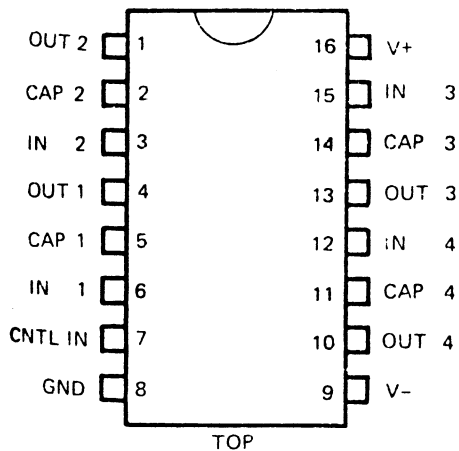
CEM3320



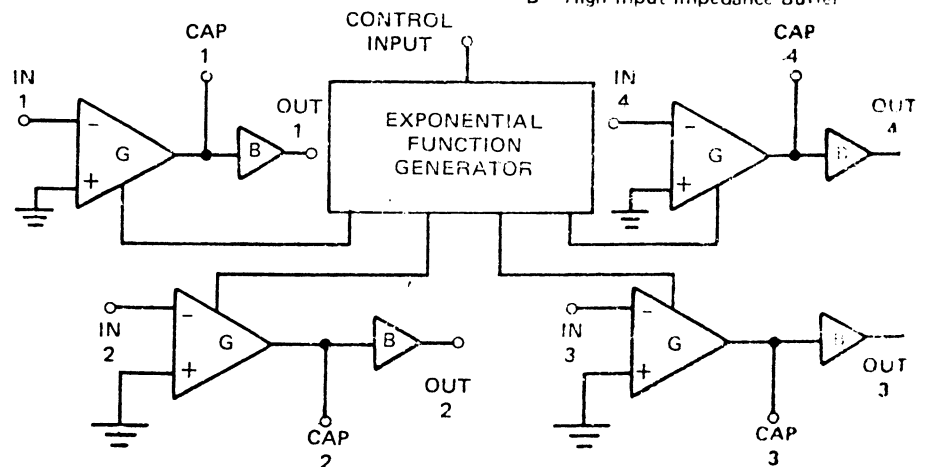
## Features

- Low Cost
- Voltage Controllable Frequency: 12 octave range minimum
- Voltage Controllable Resonance: From zero to oscillation
- Accurate Exponential Frequency Scale
- Accurate Linear Resonance Scale
- Low Control Voltage Feed-through: -45dB typical
- Filter Configurable into Low Pass, High Pass, All Pass, etc.
- Large Output: 12V.P.P. typical
- Low Noise: -86dB typical
- Low Distortion in Passband: 0.1% typical
- Low Warm Up Drift
- Configurable into Low Distortion Voltage Controlled Sine Wave Oscillator
- $\pm 15$  Volt Supplies

## SSM2040



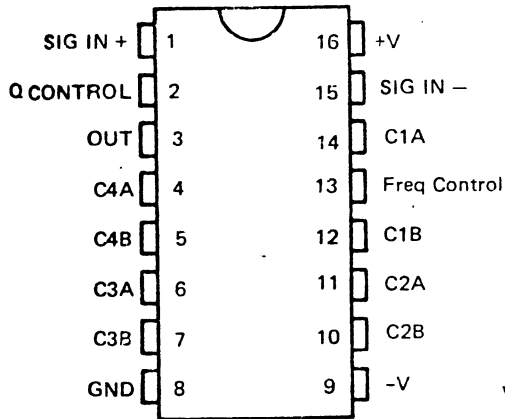
Pin Diagram



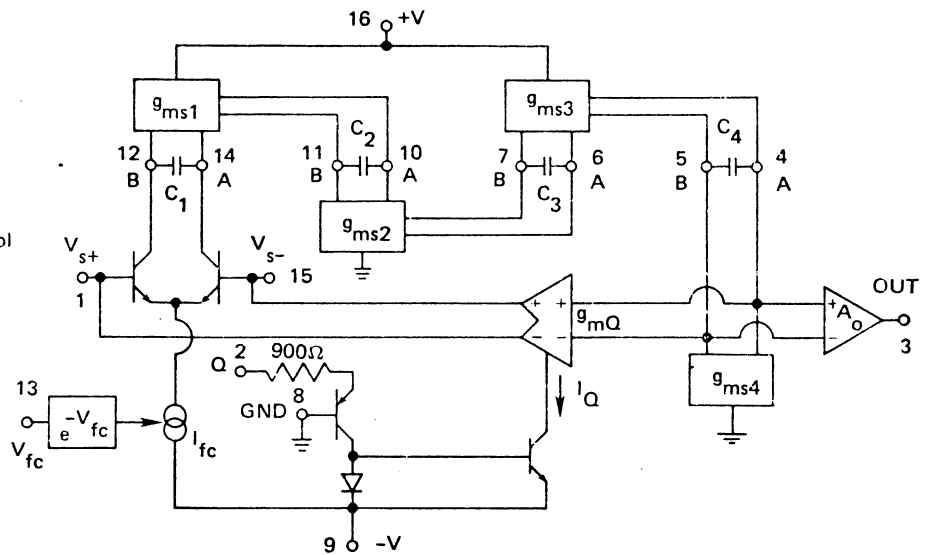
Block Diagram

G = Variable Transconductance Amplifier  
B = High Input Impedance Buffer

## SSM2044



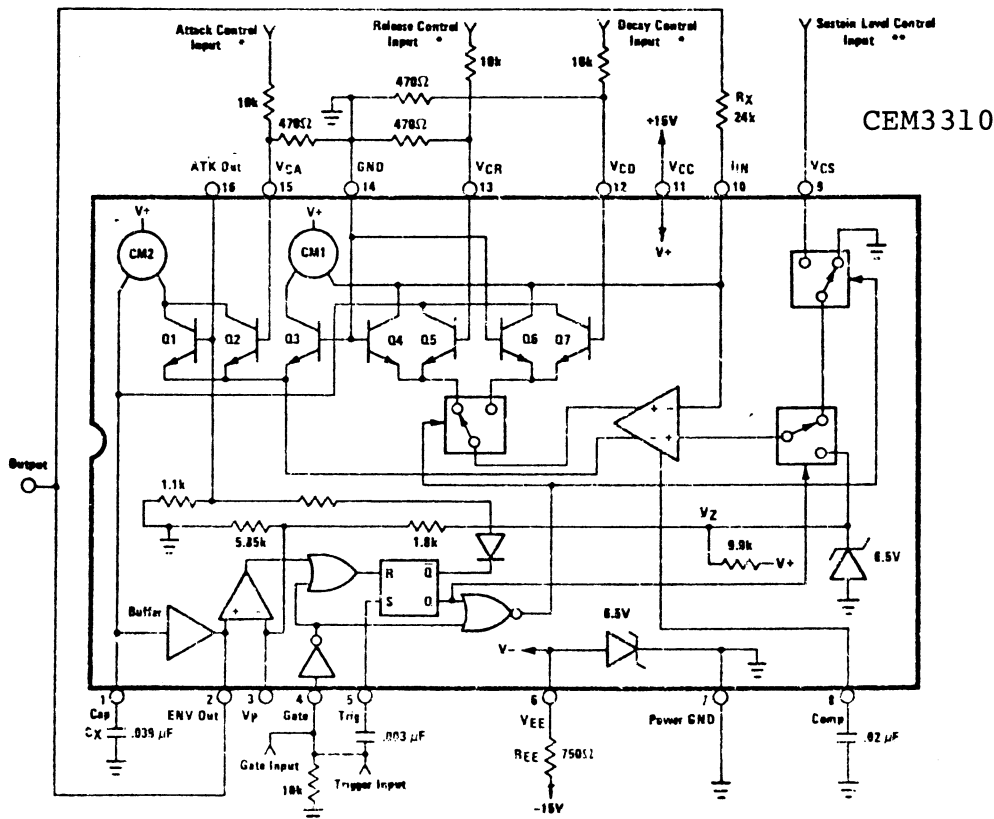
Pin Out - Top View



Functional Block Diagram

## Envelope Generators:

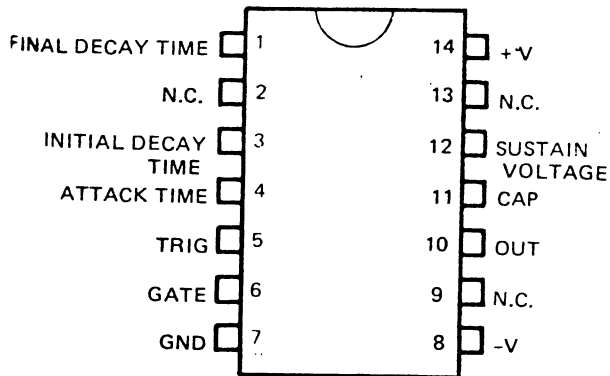
## Circuit Block and Connection Diagram



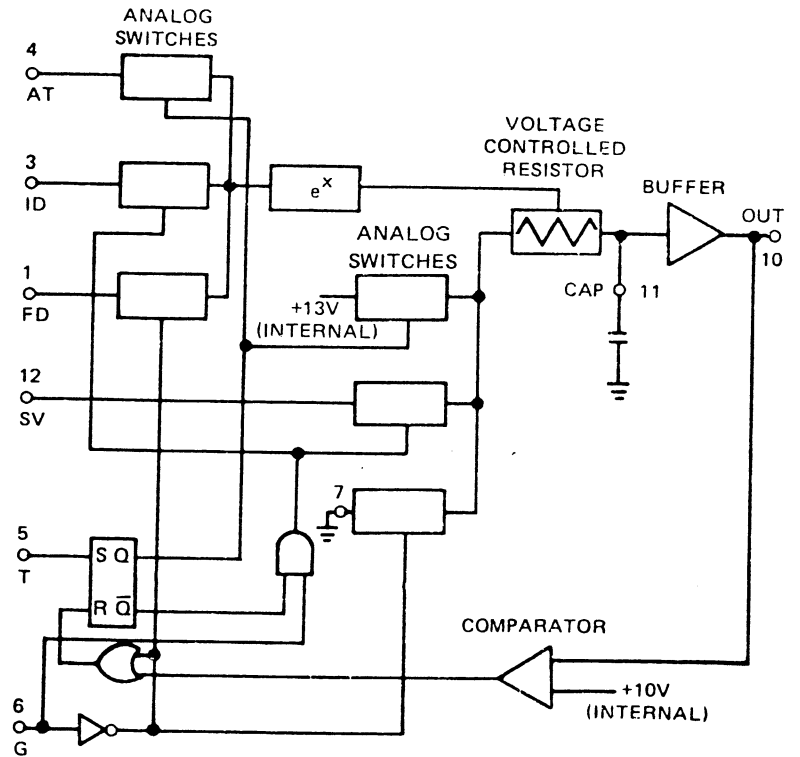
### Features

- Low Cost
- Third Generation Design
- Large Time Control Range: 50,000 min
- Full ADSR Response
- True RC Envelope Shape
- Exceptionally Low Control Voltage Feedthrough: 90μV max
- Accurate Exponential Time Control Scales
- Isolated Control Inputs
- Good Repeatability and Tracking Between Units Without External Trim
- Independent Gate and Trigger
- ± 15 Volt Supplies

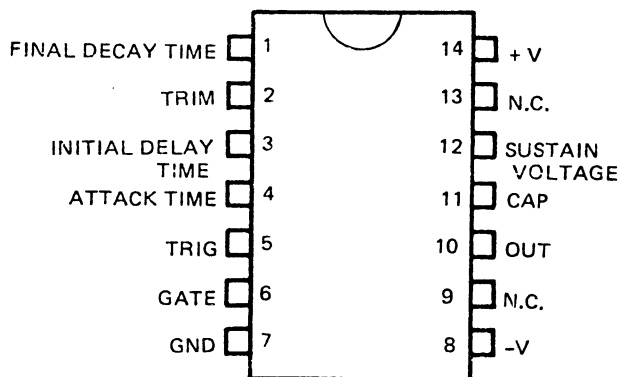
# SSM2050



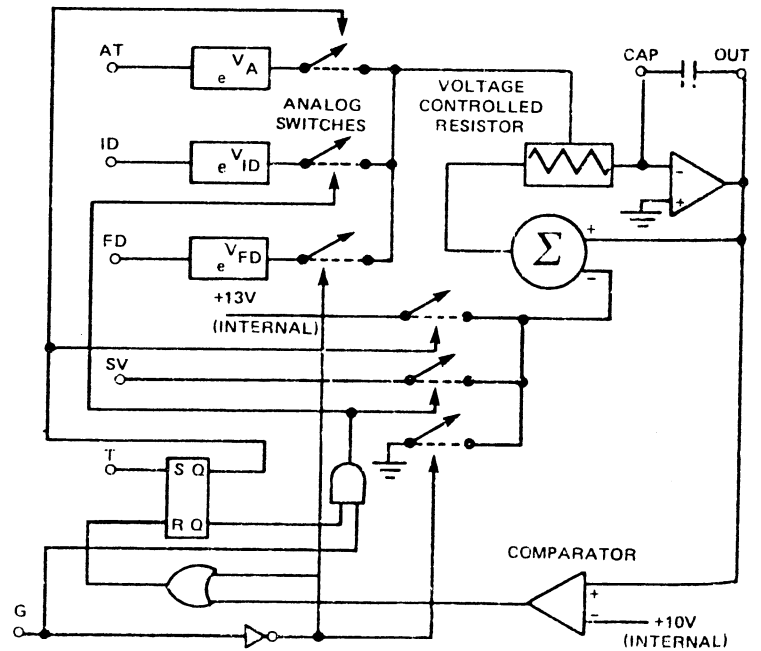
PIN DIAGRAM – TOP VIEW



# SSM2055



PIN DIAGRAM – TOP VIEW







The following listing gives some of the more useful reference materials that will help you with your building and/or use of your equipment.

▷ BOOKS:

- Hal Chamberlin Musical Applications of Microprocessors (Hayden 1980)  
Available from Micro Technology Unlimited, Box 12106, Raleigh,  
NC 27605 at a cost of \$25.00. [a better title would be:  
"Musical Electronics Including Microprocessors" as this contains  
much background and analog type information as well as digital  
and computer use]
- Don Lancaster CMOS Cookbook (Sams 1977)
- Don Lancaster Active-Filter Cookbook (Sams 1975)
- Walter Jung IC Op-Amp Cookbook (sams 1974)
- Dan Sheingold (ed) Nonlinear Circuits Handbook (Analog Devices 1976)  
Available at \$5.95 from Analog Devices, Norwood, Mass 02062.
- National Semiconductor Corp. Linear Applications Vol. 1, Linear Applications Vol. 2, Audio Handbook. These three books are very useful. Try to get them free (or pay) from your National Semiconductor distributor, or try a local Radio Shack store where they may be for sale.
- Juan Roederer Introduction to the Physics and Psychophysics of Music (Springer-Verlag, New York, 1975) You will probably have to look in a college bookstore, or have a commercial bookseller order this for you. Costs about \$7 in paperback. Extremely good introduction to how sounds are heard.

▷ PERIODICALS:

- Computer Music Journal MIT Press Journals, 28 Carleton St., Cambridge,  
MA 02142 (\$20/year)
- Synthesource, 110 Highland Ave., Los Gatos, CA 95030 (\$12.50/year,  
4 issues/year, associated with Curtis Electromusic)
- Polyphony PO Box 20305, Oklahoma City, OK 73156 \$8 per year
- Audio Amateur PO Box 576, Peterborough, NH 03458 \$14 per year

Due to the changing costs of publications, it might be best in all cases to write for the latest rates, or at least to be prepared for increases.

